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The UK's premier electronics, computing and maker magazine

Max's Cool Beans

Snake game sound effects & a VFD clock

Teach-In 12, Part 8

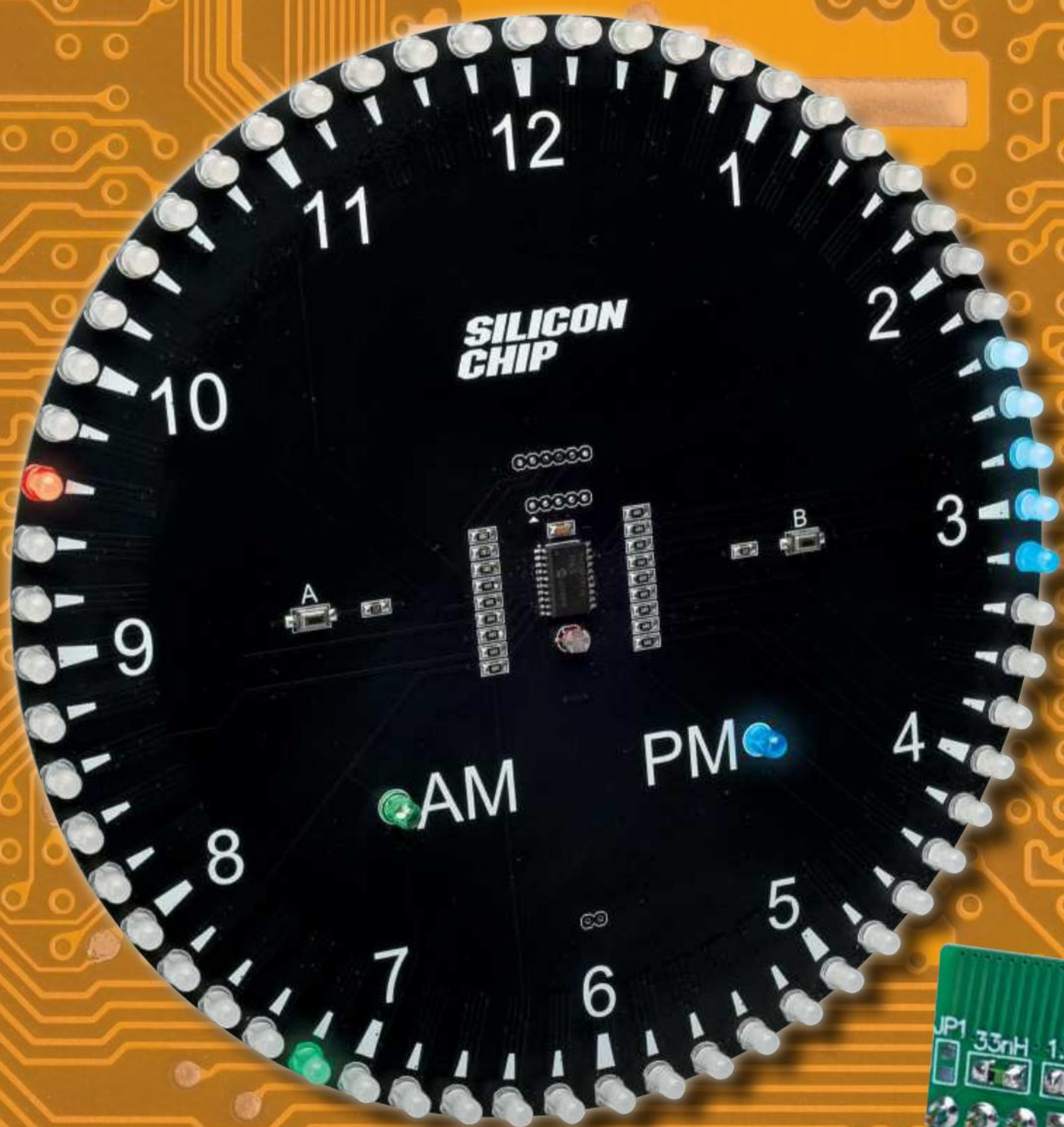
Frequency conversion with multipliers & mixers

Audio Out

An 8-way All-pass Filter PCB

Circuit Surgery

Differential and balanced signals



RGB LED 'Analog' Clock

A Simple USB Power Adaptor



DIY 433MHz Receiver



PLUS!

Power Electronics pt3 – How isolated DC/DC converters work

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Building the SSB Shortwave Receiver

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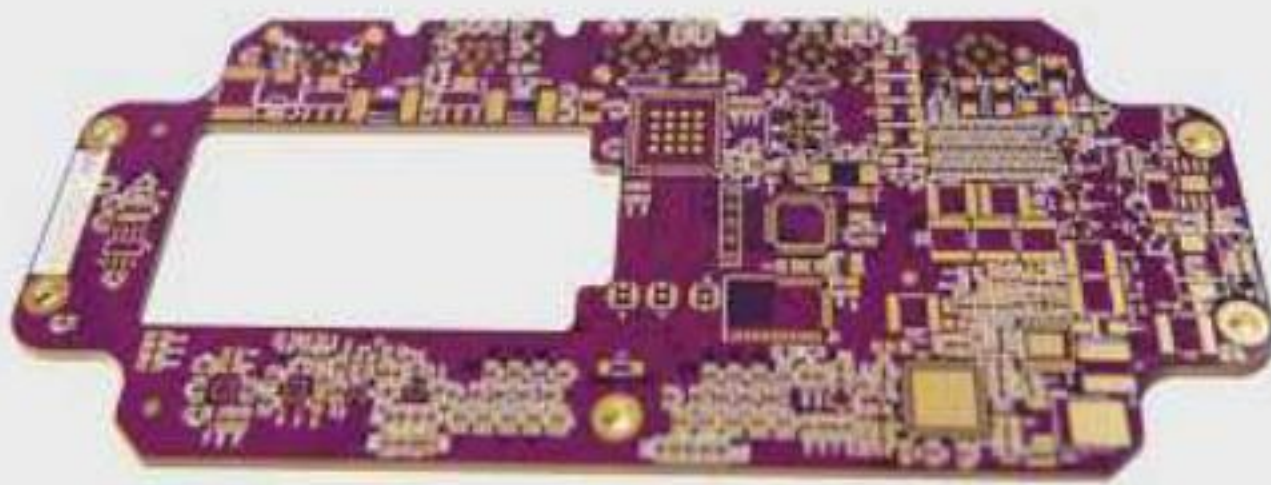


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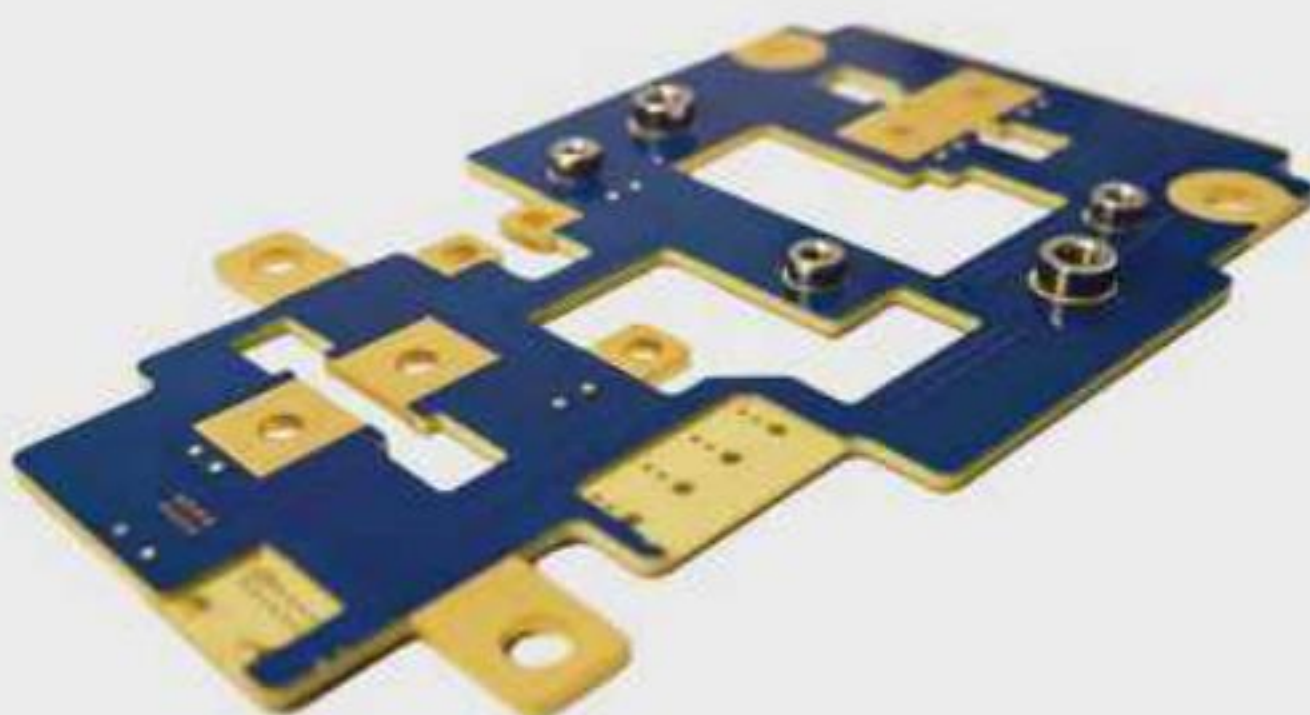


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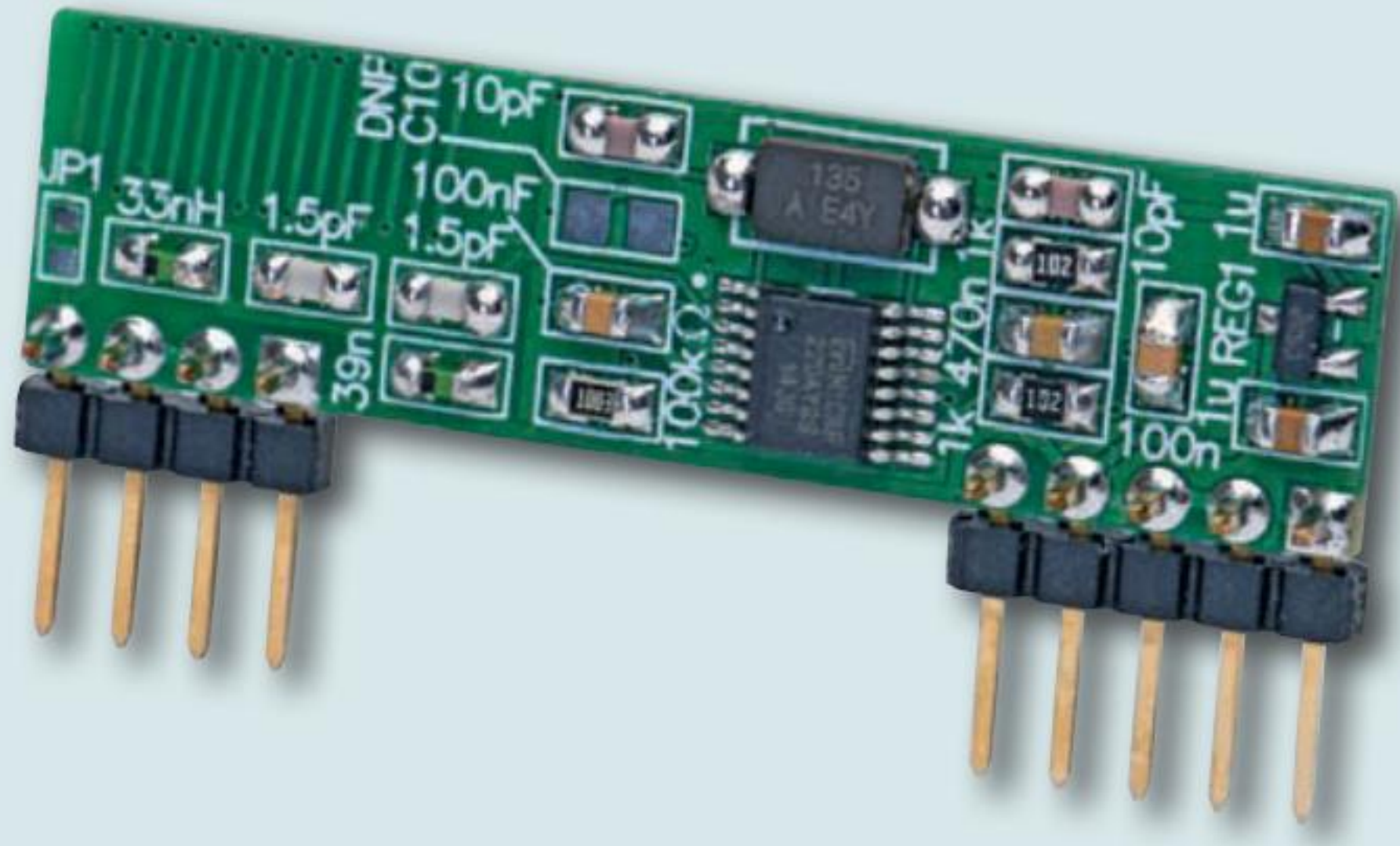
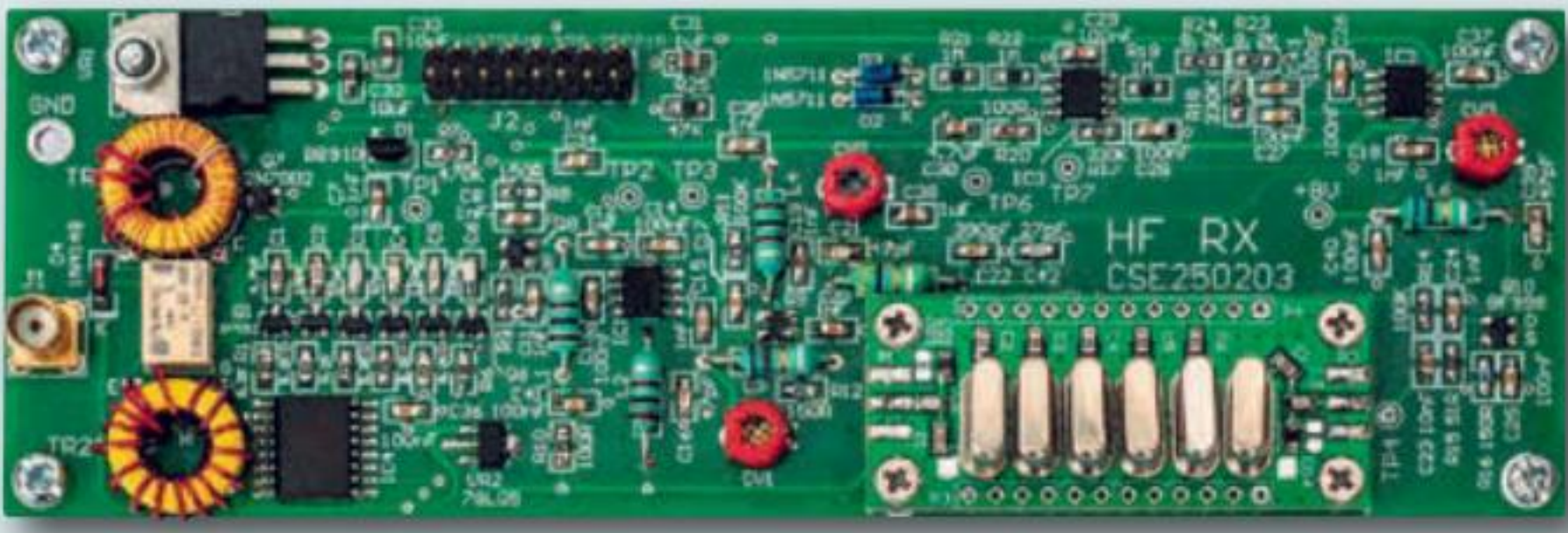
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The July 2026 issue of *Practical Electronics* will be published on Friday, 5th of June, 2026 (see p80).

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Questions about articles or projects should be sent to the editor by email: pe@pemag.au

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All reasonable precautions are taken to ensure that the advice and data given to readers is reliable. We cannot, however, guarantee it and we cannot accept legal responsibility for it.

Some projects and circuits published in *Practical Electronics* employ voltages that can be lethal. Do not build, test, modify or fix any mains-powered equipment unless you fully understand the safety aspects involved and you use an RCD (GFCI) adaptor.

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An unfortunate misprint in the May issue

Given the amount of effort involved, it's frustrating when something goes wrong with printing the magazine, especially when it is out of our control. Page 69 last month was accidentally printed with the wrong content, meaning page 67 was duplicated. The correct content is opposite. You can also download that page to print from https://www.pemag.au/PEMay26_p69.pdf – sorry about that.

It isn't "sideloading"; it's the normal way to install software

Installing software on a smartphone or similar device without using the 'official store' is often referred to as sideloading. That framing is misleading. For most of the history of computing, installing software without an official store was simply the normal way things worked.

However you obtained the software, provided the hardware and operating system were compatible, nobody needed to approve the process. The computer was yours, and you were free to decide what ran on it.

The idea that we should now be limited to manufacturer-approved software, installed only in manufacturer-approved ways, doesn't sit well with me. Even if this is presented as being "for our own good", to protect us from bad software, it comes at the cost of something fundamental: ownership. Giving up control of your computing device is not a small thing; it's insidious.

Once you no longer control what can run on your device, what exactly is stopping the manufacturer from:

- removing features after purchase
- revoking access to software you've already paid for
- scanning your files, including photos and documents
- reporting you to authorities if they dislike what's on your device
- deliberately reducing performance to encourage upgrades
- remotely disabling the device altogether

These aren't hypothetical concerns. Most of them have already happened at least once. If the manufacturer can do all of this to the device in your pocket, whose device is it, really?

It's bad enough that Apple confines users to a tightly controlled 'walled garden'. More concerning is that the main alternative, Android, is moving in the same direction: deliberately making so-called 'sideloading' (that is, installing software from a file) more difficult, while funnelling users toward an app store where the platform owner controls access and takes a cut of every transaction.

This isn't progress. It's a quiet erosion of ownership, and we shouldn't accept it.

Nicholas Vinen, Electron Publishing (Australia)*

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Pt. 2: Controlling DC-DC Converters

In front of these two blocks is a compensator, which we will describe in more detail below. The input to the compensator is an error, e , that is the difference between the reference setpoint v_r and the feedback signal v_{fb} . The feedback signal is derived from the output voltage we want to control by some sensor, with a gain of H .

These blocks are shown in a more familiar way in Fig.3. The modulator includes the Mosfet and diode switches, plus the circuitry that converts a control voltage to pulse-width-modulation to drive the Mosfet. At its simplest, this is just a comparator with an output that is high whenever the control voltage exceeds a ramp waveform at the switching frequency.

The modulator therefore takes a control voltage, v_c , and produces the average voltage, $\langle v_x \rangle$. The modulator has a gain of G_m and it is easy to see that this must be the unitless quantity $G_m = V_{SRC}/V_{RAMP}$. If we assume the ramp voltage is 1V, the modulator gain simplifies to the source voltage, which is 12 in our example.

It is worth noting that there is no need for any switching to occur in this model of the modulator. The modulator produces an output equal to the average voltage over one switching period. This voltage may vary with time, but the time scale we are concerned about is much longer than the $2\mu s$ switching period.

Because this modulator controls the average voltage applied to the filter, this converter is said to be employing voltage mode control. There is another way to do it, which we will come to later.

The sensor gain is trivial to calculate. It is simply the divider ratio represented by the two resistors, so $H_s = R1 \div (R1 + R2)$, or 0.204 in this case. The buffer is there to isolate the compensator from the divider.

We will ignore the compensator for the moment, but notice that I have absorbed the error summing junction into the black box and added a positive input for the reference and a

Fig.2: this classic depiction of the DC-DC converter closed-loop system is useful to understand the circuit blocks we need to model.

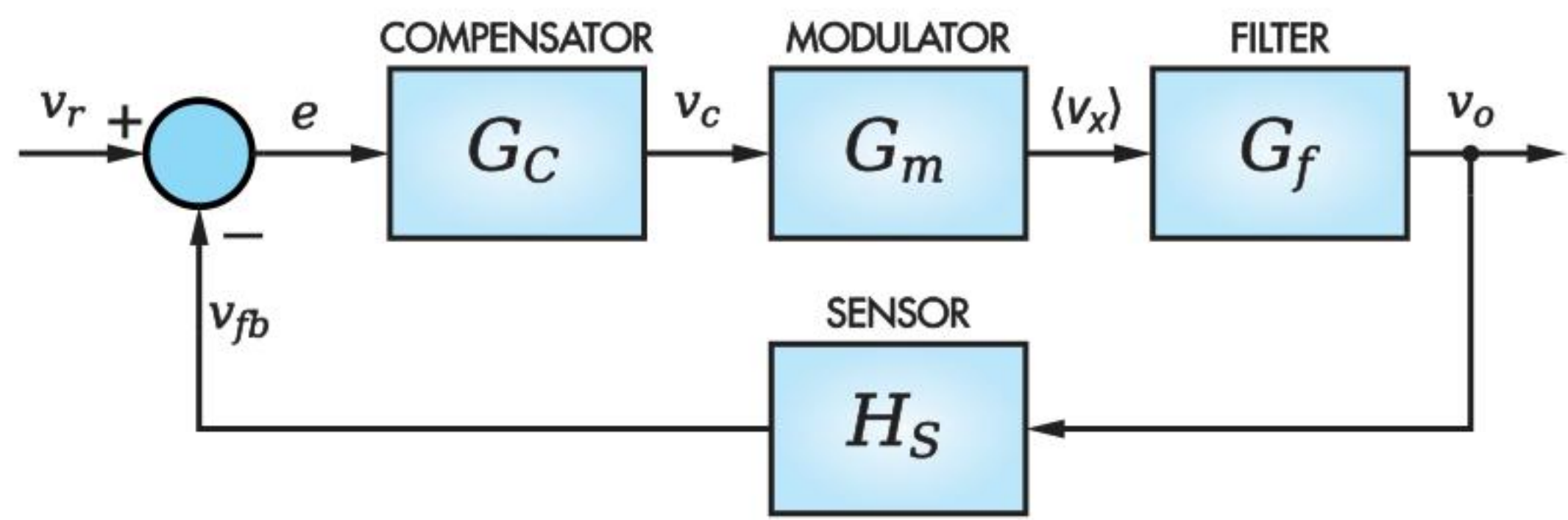
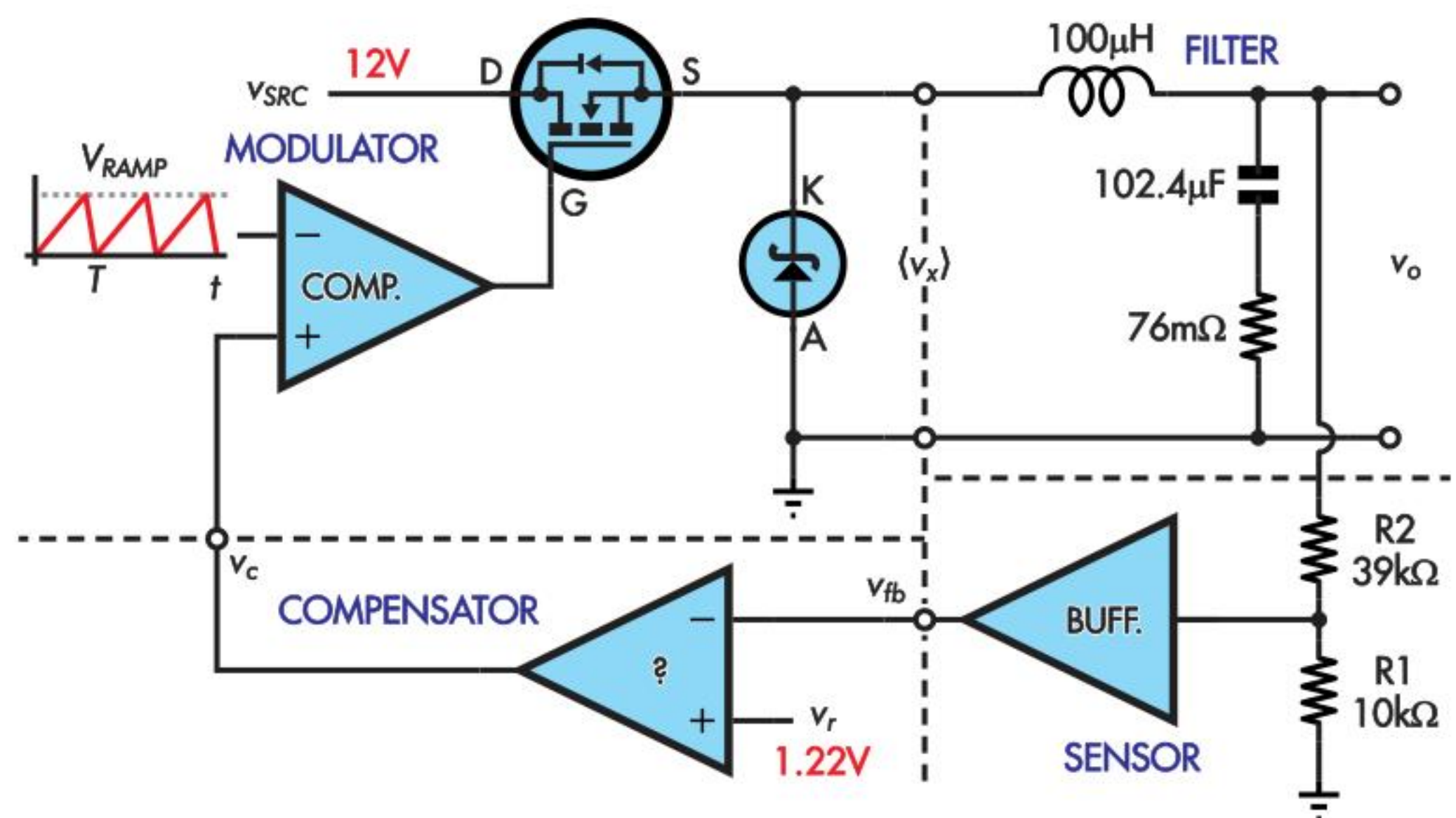


Fig.3: the DC-DC converter arranged according to the block diagram in Fig.2 is the starting point for the development of a control model of the converter. The compensator is the missing piece of the puzzle at this stage.



negative input for the feedback. This anticipates that the compensator will probably involve op amps.

Simplifying the complex-frequency domain

Describing the filter and compensator gains requires us to dip our toes into something called the complex-frequency domain, also known as the s-domain. Analysis of the behaviour of capacitors and inductors involves differential equations, which we usually describe in the time domain – for example, the relationship between voltage and current in an inductor is $v = L di/dt$.

We can describe the same equations in the s-domain, which makes the mathematics much simpler, because differentiation is reduced to multiplication and integration is reduced to division. The conversion from the time domain to the s-domain requires something called a Laplace transform.

You don't need to know how to do this (unless you are an electrical engineering student) because you can write the s-domain expression for most common components very easily without having to do any maths.

The “s” in s-domain is a complex frequency, which is just a way of describing a unit-amplitude sinewave with a given frequency and phase. We won't, however, have to deal with any complex numbers in this explanation – we can simply treat s like any other algebraic variable.

To transform the inductor mentioned above to the s-domain, the time differential (d/dt) is simply replaced by s to get the s-domain equivalent relationship $v = i \cdot L \cdot s$, which I hope you can see is kind of analogous to Ohm's law. The complex impedance of the inductor, indicated by $Z(s)$, is therefore simply Ls .

Using a similar logic, we can arrive at the s-domain expressions for the voltage/current relationships of common parts, and hence their complex impedances, shown in Table 1.

To show just how useful this transformation is, we will use these complex impedances to calculate the transfer function of a simple RC low-pass filter shown at the top of Fig.4. If you substitute in the complex impedances from the table and use the normal voltage divider equation, you get the s-domain transfer function $1 \div (RCs + 1)$ after a little bit of algebraic manipulation. It's that easy.

Poles and zeros

You can see that there must be a particular value of complex frequency s that makes the denominator of the

Table 1 – passive component complex impedances

Component	Time Domain	s-Domain	Complex Impedance $Z(s)$
Inductor	$v = L di/dt$	$v = iLs$	Ls
Capacitor	$i = C dv/dt$	$i = vCs$	$1/Cs$
Resistor	$v = Ri$	$v = iR$	R

Teach-In 2026

by Mike Tooley



World of Wireless – An Introduction to Radio and Wireless Technology, part 12.8

Frequency conversion with multipliers and mixers

Last month, we explored some useful techniques for analysing RF networks and devices, including S-parameters, vector network analysis (VNA), and spectrum analysis (SA).

VNA plays a vital role in RF engineering by offering precise measurements of network and device properties across a wide frequency spectrum. Spectrum analysis is essential for examining the components within RF signals to evaluate spectral purity and identify spurious emissions.

This month, our focus shifts to frequency conversion methods used in receivers, transmitters and synthesisers, with particular emphasis on how multipliers and mixers can change the frequency of a signal.

A comprehensive understanding of circuit operation can be achieved

through experiments conducted with software simulation. Accordingly, this month we will be introducing the use of SPICE software to investigate multiplier and mixer circuits. Our *Hands-On* project is a 50MHz Digital Frequency Meter that's ideal for making a variety of RF measurements.

Frequency conversion

In radio and wireless, the ability to change or 'convert' the frequency of a signal is essential to the operation of all but the most basic types of receivers and transmitters. Frequency changers, such as multipliers and mixers, are the essential devices that make this possible.

By shifting signals to higher or lower frequencies, they allow for easier filtering, improved selectivity, and enhanced system performance.

Frequency conversion is fundamental to superheterodyne receivers (see Part 2, December 2025), where incoming signals are converted to a fixed intermediate frequency (IF). This permits more effective filtering, processing and demodulation.

In transmitters, generated signals can be moved to frequencies suitable for transmission in a desired band. Understanding the principles and applications of frequency changers is therefore vital for anyone involved in RF design and wireless communications.

Fig.8.1 shows the basic principles of the two types of frequency changer. In Fig.8.1(a) the input frequency, f_{in} , is multiplied by a factor of n to produce an output frequency, f_{out} , where $f_{out} = n \times f_{in}$. Typical values of n range from two to five.

For example, an arrangement

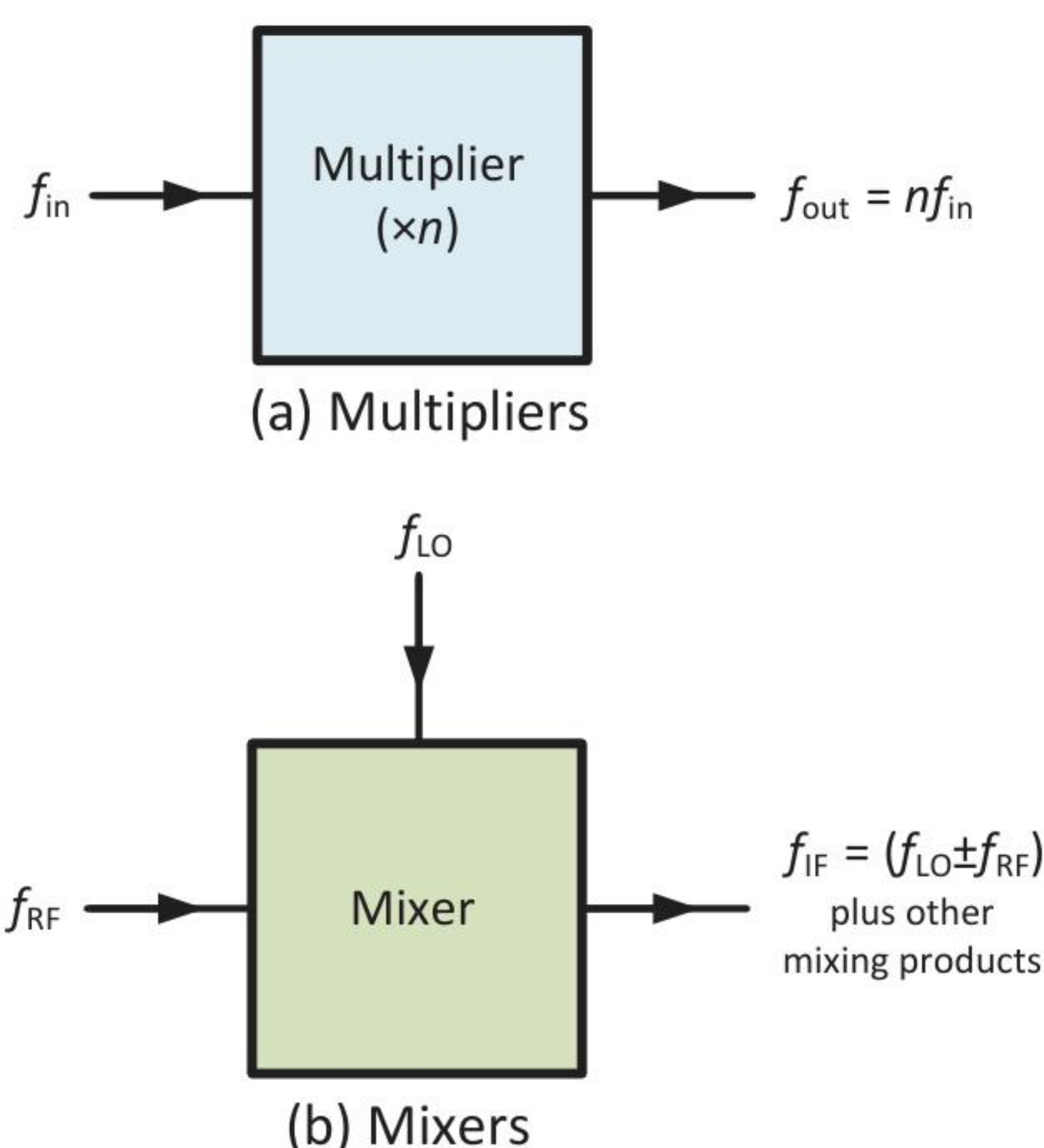


Fig.8.1: the difference between a multiplier and a mixer.

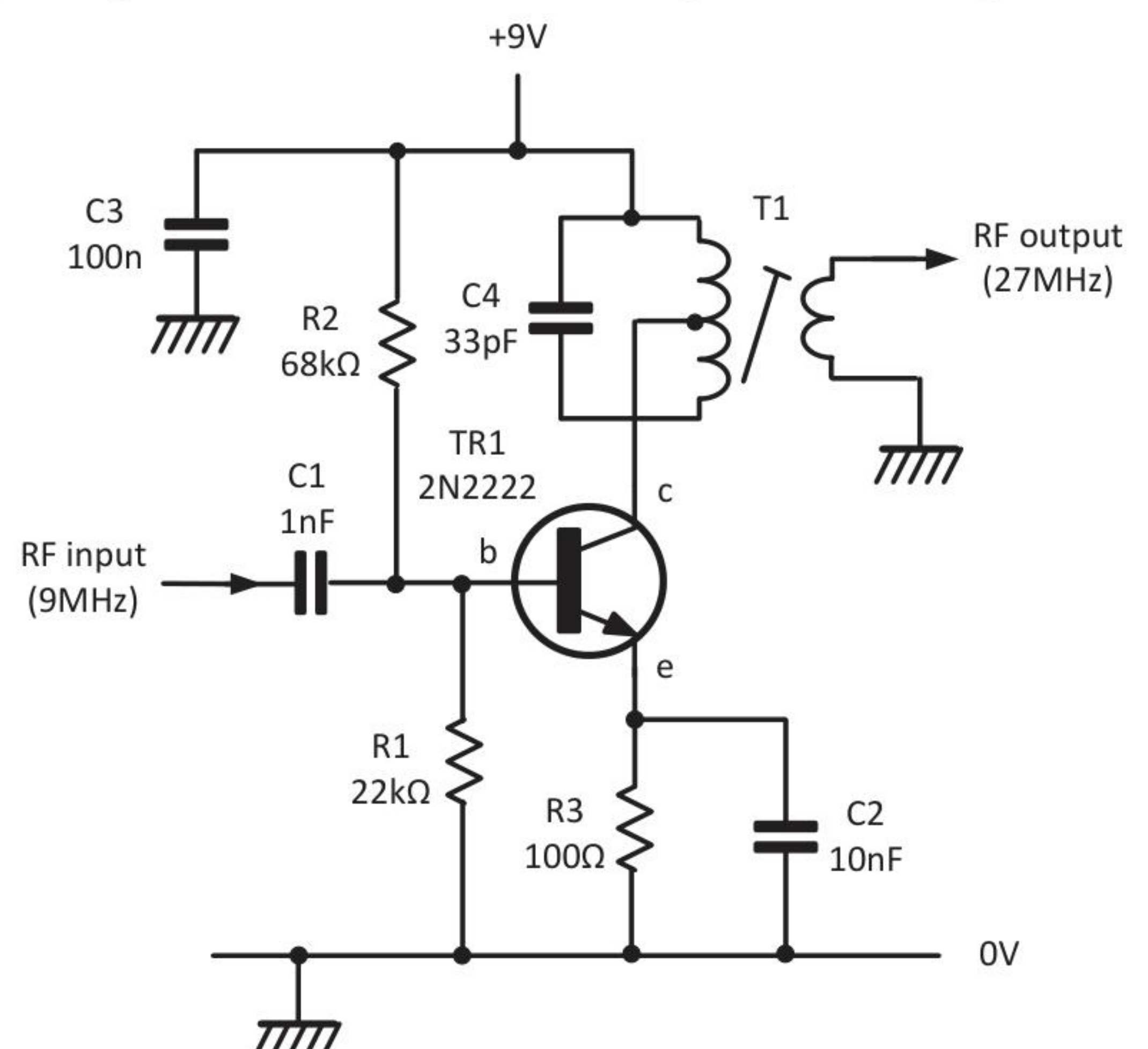


Fig.8.2: the circuit of a simple frequency multiplier.

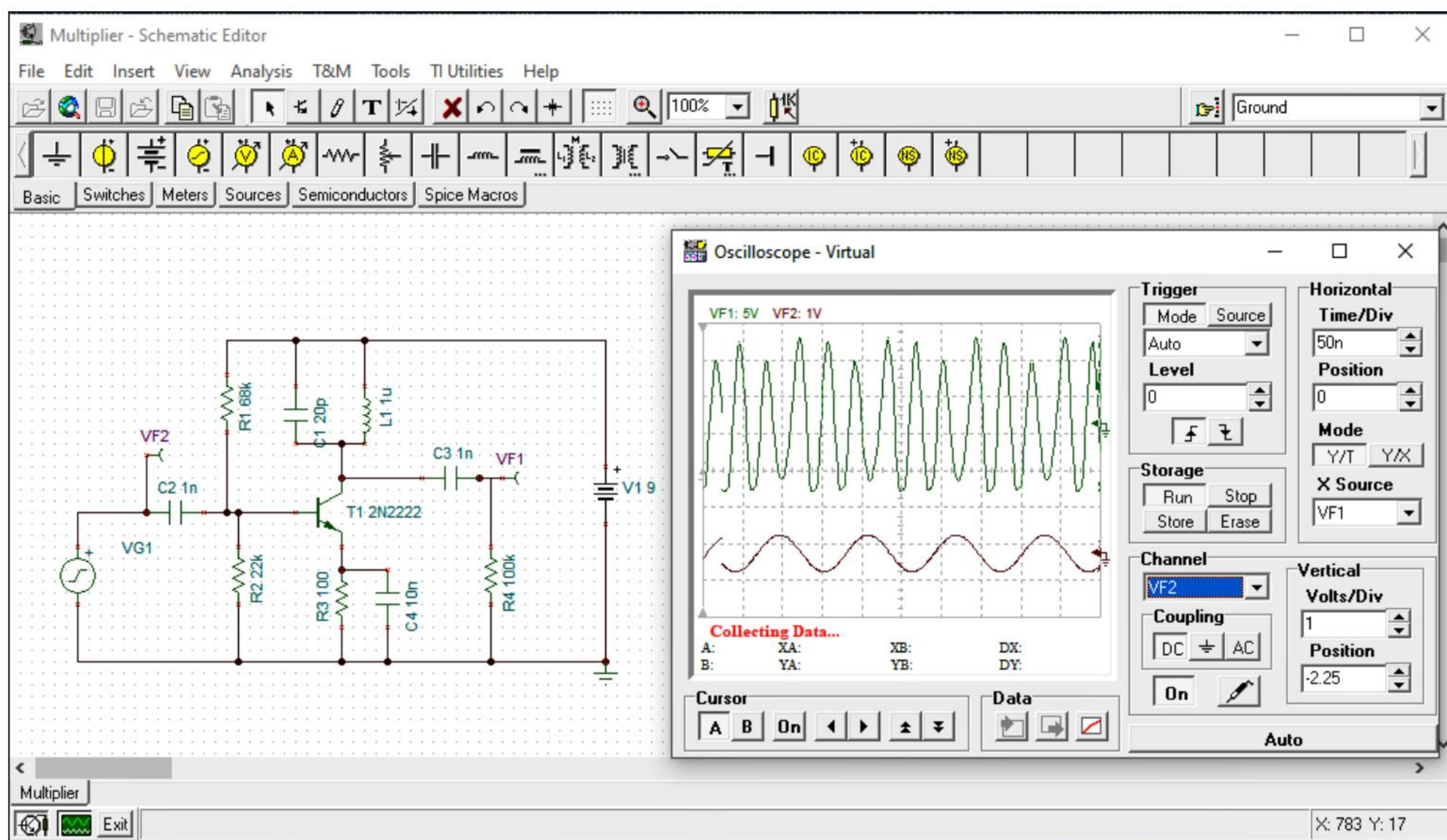


Fig.8.3: a Tina-TI simulation of the circuit in Fig.8.2.

where $n = 2$ (known as a ‘doubler’) will produce a signal at 50MHz when supplied with an input at 25MHz. In a similar manner, a circuit where $n = 3$ (known as a ‘tripler’) will produce a 21MHz signal when supplied with an input signal at 7MHz.

In Fig.8.1, the intermediate frequency output, f_{IF} , is generated by mixing the input frequency, f_{RF} , with the local oscillator (LO) frequency, f_{LO} . The mixing process is not simply additive as it is with a basic audio mixer, but instead relies on non-linear characteristics within the mixer.

In additive mixing, two signals are combined by simply summing their voltages. The resulting output will contain both original frequencies, but no new frequency components are generated. By contrast, multiplicative mixing uses a non-linear device (such as a diode or transistor) to produce additional frequency components, as well as the sum and difference of the original frequencies.

These multiplicative ‘mixer products’ generate signals at new frequencies, thereby allowing signals to be shifted to different frequency bands.

In Fig.8.1(b), the intermediate frequency at the output of the mixer can be calculated as $f_{IF} = (f_{LO} \pm f_{RF})$. So, for example, if an IF of 455kHz is to be obtained from an RF input at 14MHz, the local oscillator would need to be running at either 14.455MHz (above the RF input) or 13.545MHz (below the RF input).

Either of these two local oscillator frequencies will produce the required intermediate frequency, but there can be reasons to prefer one or the other.

Popular intermediate frequencies are 455kHz and 470kHz (AM), or 1.6MHz, 9MHz, 10.7MHz and 21.4MHz (FM); the value chosen depends on the application, modulation method, and receiver complexity. It’s worth noting that some receivers use double and even triple-conversion techniques, in which case several mixers will be present.

Multipliers

A basic form of frequency multiplier is shown in Fig.8.2. This circuit uses a conventional NPN transistor connected in common-emitter mode. The collector load is a tuned circuit formed by the primary winding of T1 and capacitor C4. This parallel-tuned circuit is resonant at 27MHz (three times the RF input frequency).

To provide a better match and improve the Q-factor of the load, the collector of TR1 is connected to a tap in T1’s primary at about its midpoint. When a 9MHz input is applied to the base of TR1, the output voltage from the secondary of T1 will be reasonably sinusoidal at three times the input frequency (the circuit thus acts as a ‘tripler’).

Fig.8.3 shows the Tina-TI simulation of Fig.8.2. For simplicity, we’ve replaced the tuned transformer with a simple LC tuned circuit (see Part

2 of Teach-In). The values of C1 and L1 (20pF and 1μH, respectively) are chosen so that the tuned circuit is resonant at approximately 27MHz.

In Fig.8.3, we’ve adjusted the signal source, VG1, so that it produces a sine wave at 9MHz with an amplitude of 500mV. The resulting rectangular pulses of current at the collector of TR1 excite harmonic resonance in the collector tuned circuit, so the collector voltage will be a sine wave at three times the input frequency.

In Fig.8.3, we’ve used Tina’s excellent virtual oscilloscope to display the input and output waveforms. Note that the input (shown in red) has a peak-to-peak value of about 1V, whereas the output (shown in green) is almost 20V peak-to-peak.

Notice that some of the input frequency signal is present in the output waveform, but at a significantly lower level. The residual amplitude can be further reduced by increasing the Q-factor of the load, as in the transformer-coupled arrangement shown earlier in Fig.8.2.

If you would like to experiment with the component values in Fig.8.3, it can be instructive to calculate values for L1 and C1 to make the circuit act first as a doubler ($n = 2$) and then as a quadrupler ($n = 4$). You can calculate the required values using the formula shown on page 42 in Teach-In Part 2: introducing RF components and circuits.

Simply click on the components (L and C) in the simulation to change

their values and observe the effect on the output waveform. If you don't already have this excellent simulation tool, Tina-TI can be freely downloaded from <https://www.ti.com/tool/TINA-TI>

Mixers

Fig.8.4 shows a basic mixer. The RF input is applied to the base of TR1 via capacitor C1, while the local oscillator is coupled into the emitter by capacitor C2. The intermediate frequency output is extracted from the collector of TR1; C4 and the primary of T1 are resonant at the intermediate frequency (in this case, 450kHz).

To remove the unwanted mixing products (eg, those resulting from signal components just above and below 3.5MHz) a filter is connected at the output of T1. This passive component accepts the intermediate frequency but rejects the other mixing products (of which there are many).

IF filters can be based on LC networks, quartz crystals or low-cost ceramic resonators. Typical filter bandwidths are 15kHz for AM broadcasts, 150kHz for FM broadcasts or 9kHz for AM voice communication. Narrow-band modes, such as single-sideband (SSB), continuous wave (CW), and low-speed data (such as PSK31) all benefit from narrower filters with bandwidths of 3kHz or less.

Fig.8.5 shows the Tina-TI simulation of Fig.8.4. For simplicity, we've once again replaced the tuned

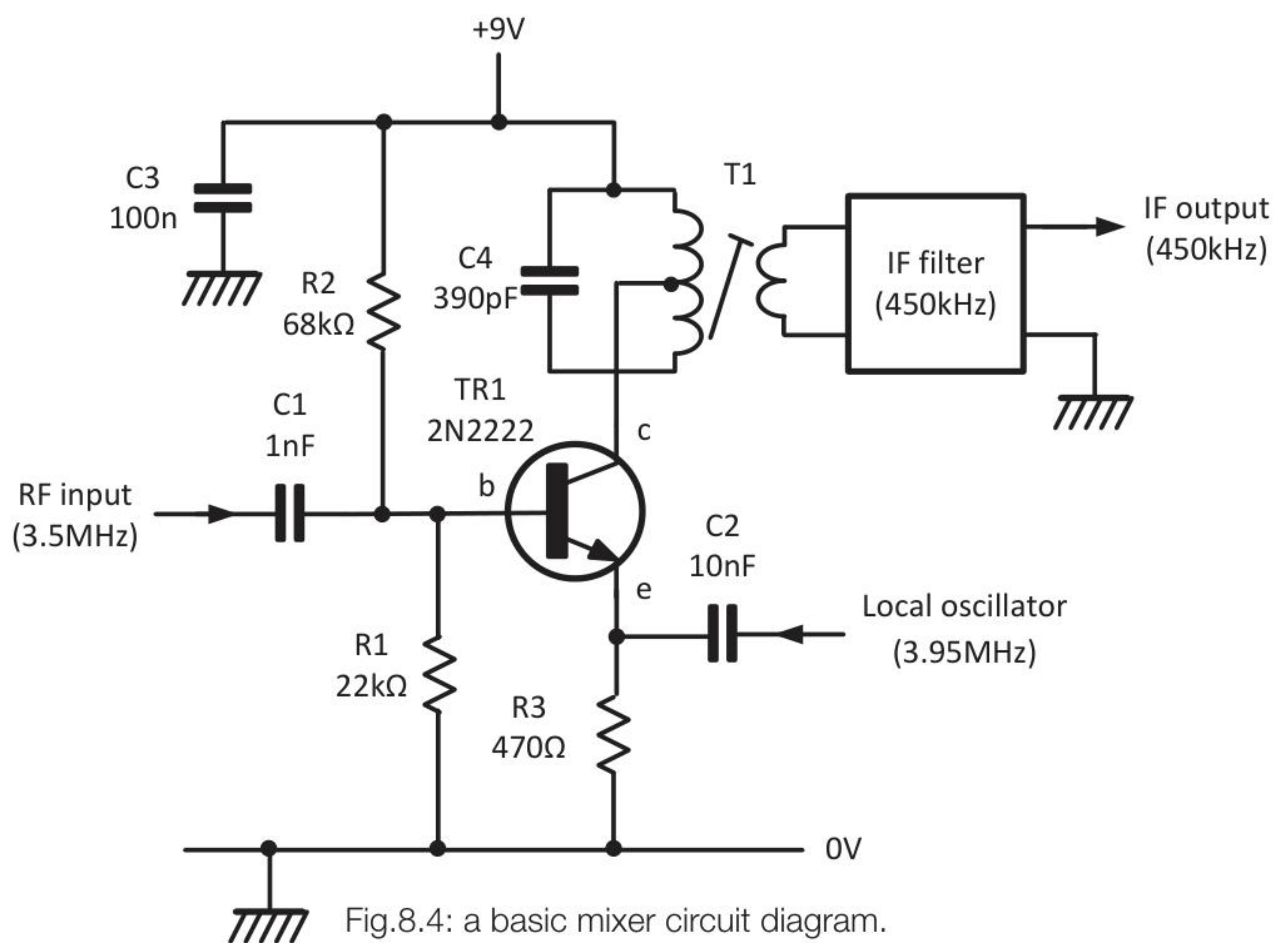


Fig.8.4: a basic mixer circuit diagram.

transformer with a simple LC tuned circuit. The values of C3 and L1 (390pF and 330μH, respectively) are chosen so that the tuned circuit is resonant at approximately 450kHz.

We've once again used Tina's virtual oscilloscope to display the input and output waveforms. The RF signal source, VG1, should be adjusted so that it produces a sinewave at 3.5MHz with an amplitude of 500mV. To produce a 500kHz IF, the local oscillator source (VG2) needs to be set to 4MHz with an amplitude of 500mV.

Note how the signal input (shown in red) has a peak-to-peak value of

about 1V, whereas the output (shown in green) is once again around 20V peak-to-peak (indicating that there is some gain present in the mixer).

It's also worth noting that some of the local oscillator signal is present in the output waveform, but at a much lower level. The residual LO signal content can be further reduced by increasing the Q-factor of the load, as in the more selective transformer-coupled arrangement shown earlier in Fig.8.4.

You might like to experiment with different RF signal input and local oscillator frequencies in the settings

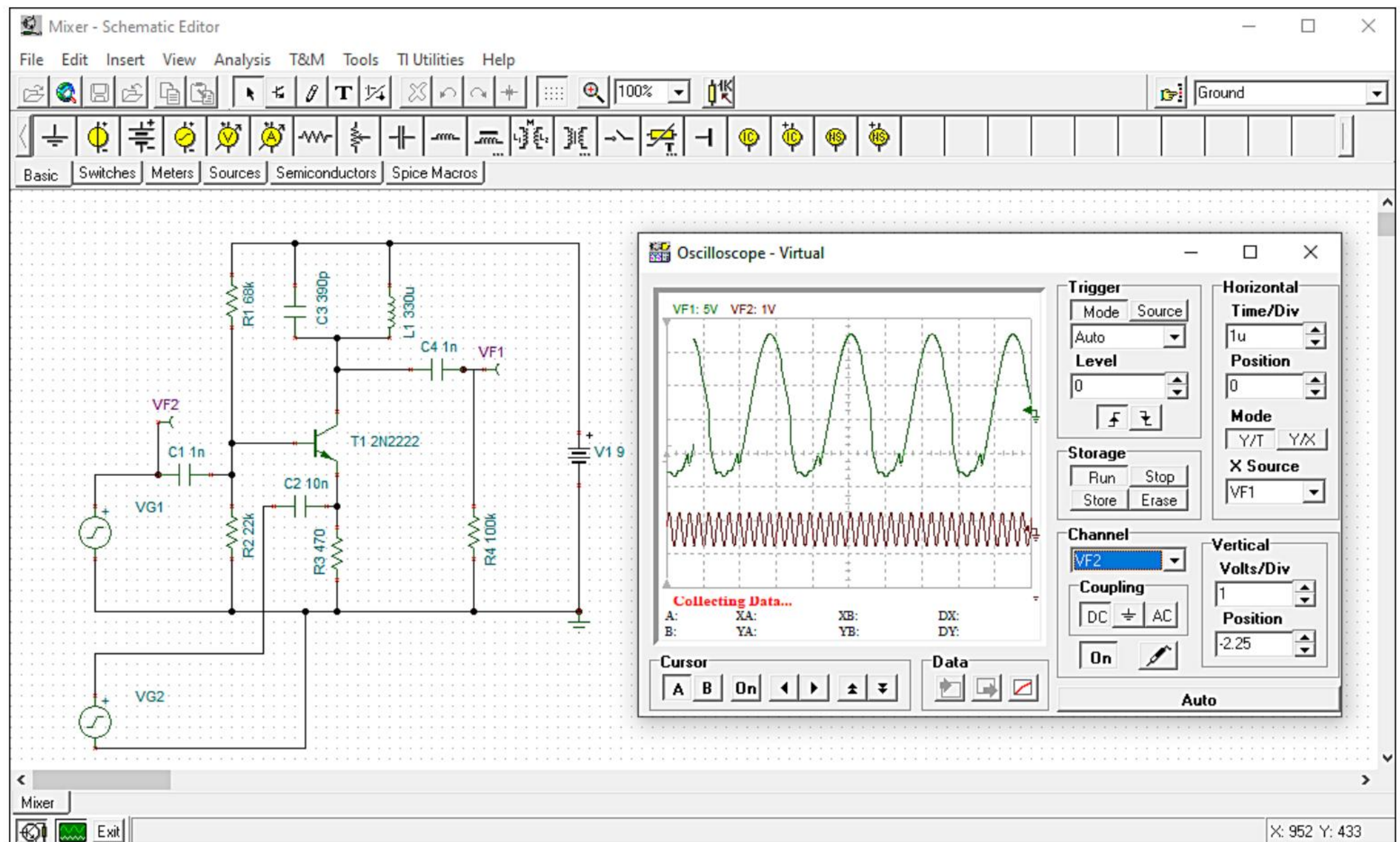


Fig.8.5: a Tina-TI simulation of the circuit in Fig.8.4.

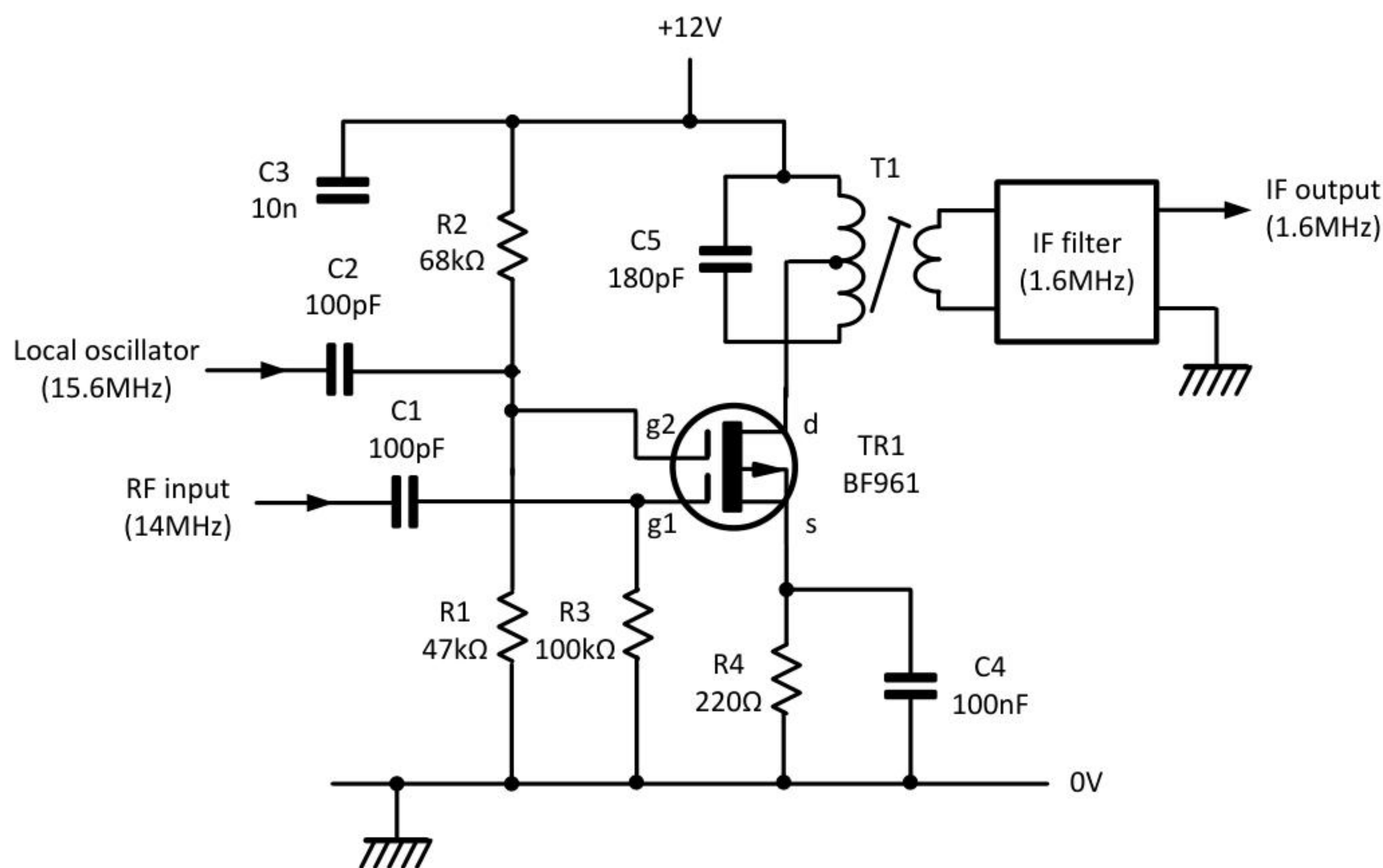


Fig.8.6: an improved mixer based on a BF961 dual-gate Mosfet.

of the two voltage generators, VG1 and VG2. For example, it can be instructive to see if you can produce the same intermediate frequency with the local oscillator injection on the low side of the RF input (ie, set VG2 to 3MHz rather than 4MHz).

You could also try a higher intermediate frequency of, say, 1.6MHz, with the local oscillator (VG2) set to 5.1MHz.

Fig.8.6 shows the circuit of an improved mixer arrangement based on a dual-gate Mosfet. TR1 is connected in common-source mode with the RF signal input applied to gate 1 (G1) and the local oscillator applied to gate 2 (G2).

Since both gates have a very high impedance, the circuit requires very little signal and local oscillator drive

power. The values shown in Fig.8.5 are for operation at 14MHz with an IF at 1.6MHz. High-side local oscillator injection is provided at 15.6MHz.

A practical implementation of Fig.8.6 is shown in Fig.8.7. With RF input and local oscillator frequencies of 7.2MHz and 7.655MHz respectively, this arrangement uses a low-cost ceramic filter to produce an IF at 455kHz.

The LT455EU filter provides a 6dB bandwidth of $\pm 7.5\text{kHz}$ and is thus ideal for AM broadcast reception. Additional filters in the LT455 series offer bandwidths as low as $\pm 2\text{kHz}$, making them a better choice for SSB reception.

Image channel rejection

As previously mentioned, a

mixer produces the same IF when it receives signals at two distinct RF input frequencies: one above and one below the local oscillator frequency. Of these, one represents the desired signal, while the other is unwanted and referred to as the 'image channel'. The frequency difference between the desired signal and the image channel is twice the intermediate frequency.

Image channel rejection in a superhet receiver is the ability of a receiver to suppress or reject signals that appear at the image frequency. Image channel rejection is usually specified in dB and is the ratio of the amplitudes of the wanted and image channel signals that produce the same IF output amplitude.

Effective image channel rejection ensures that unwanted signals do not interfere with the reception of wanted signals, improving image channel selectivity and overall receiver performance.

To mitigate the impact of signals present in the image channel, one or more stages of input filtering are essential. Increasing the intermediate frequency reduces the filtering requirements for the RF input stages. Consequently, a receiver can achieve significantly higher image channel rejection with an IF of 1.6MHz than with, for example, 455kHz.

Typical values of image rejection range from around 40dB for an inexpensive AM superhet receiver to better than 70dB for a high-end communications receiver featuring dual conversion and two intermediate frequencies.

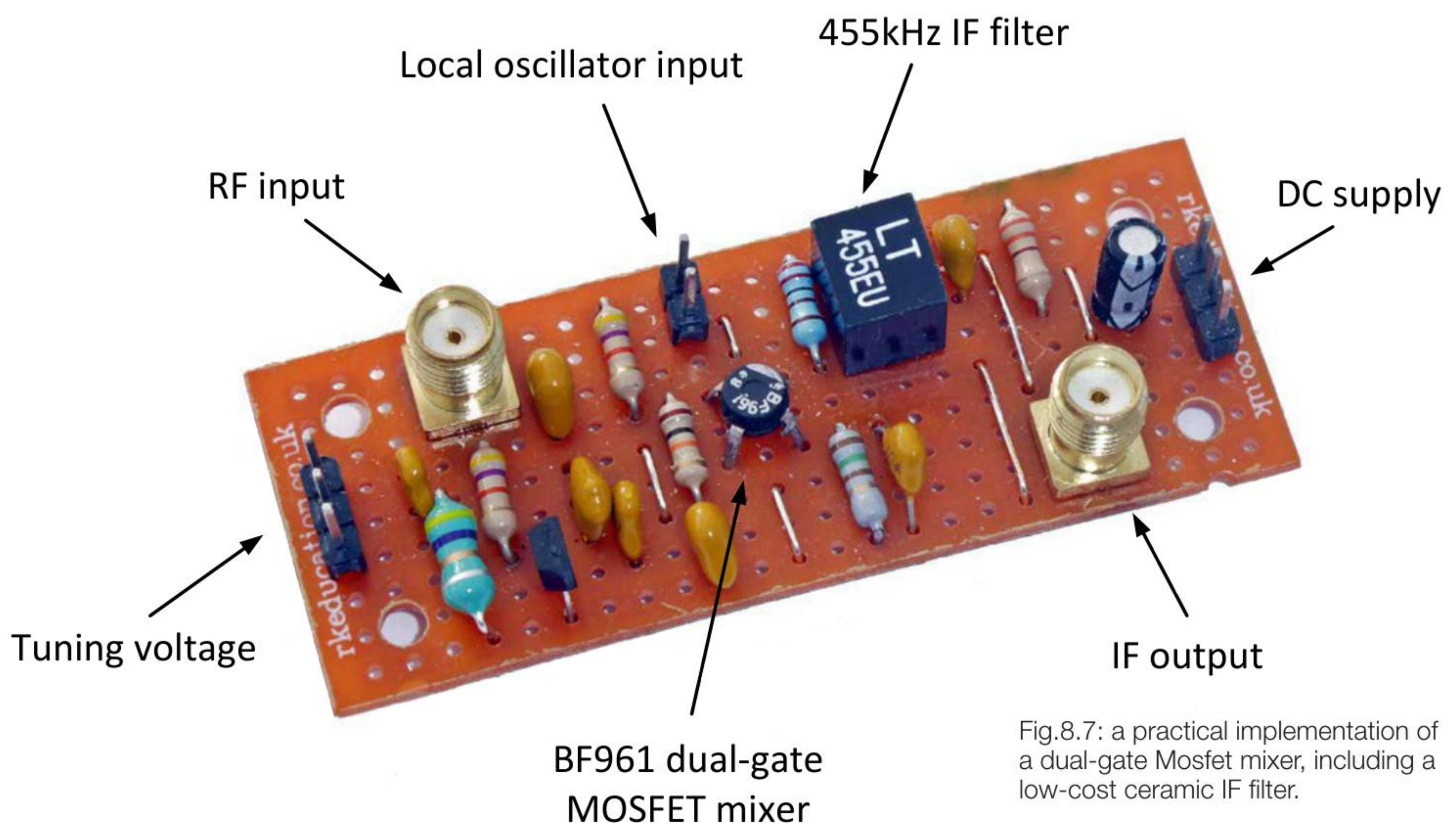
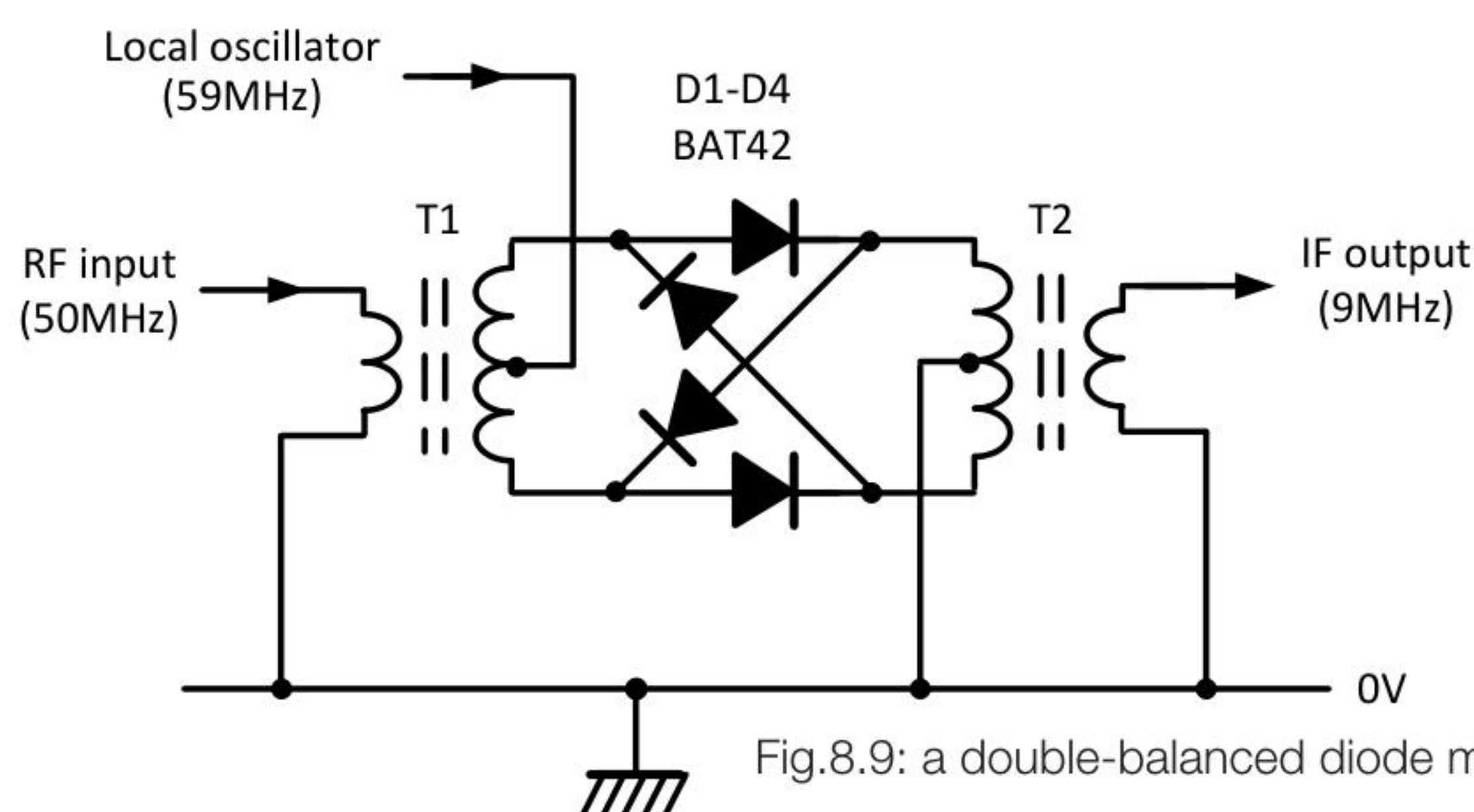
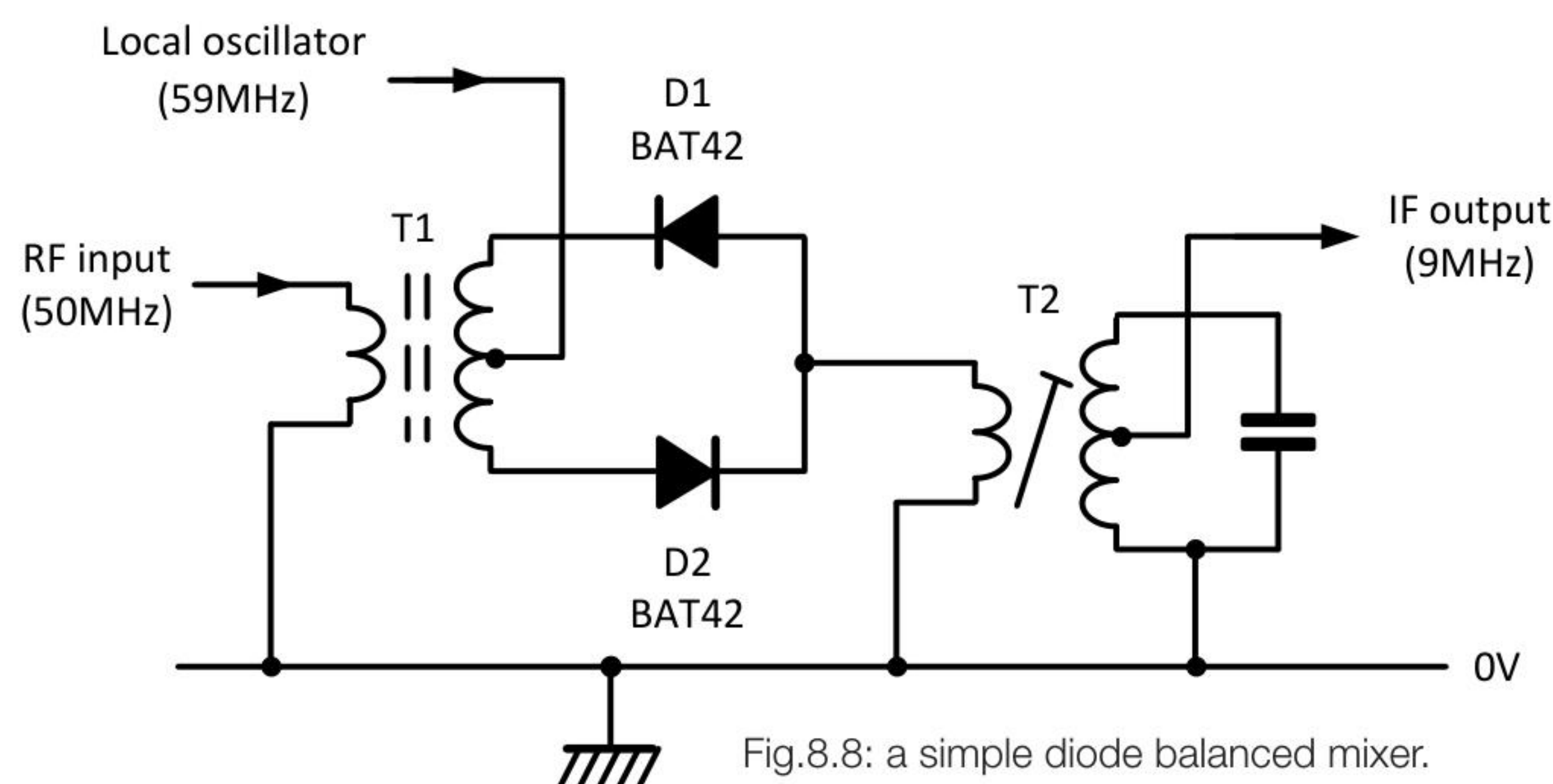


Fig.8.7: a practical implementation of a dual-gate Mosfet mixer, including a low-cost ceramic IF filter.



It is often easier to obtain the best image rejection when the LO is above the received frequency rather than below it because that shifts the image frequency above the broadcast band, in the case of an AM receiver, allowing a low-pass filter to reject it without affecting desired signals.

Conversion gain

Many mixers provide a small (but useful) amount of gain. This 'conversion gain' is simply the ratio of the IF output signal voltage amplitude to the RF input signal voltage. Conversion gain is usually expressed in dB, and typical values of conversion gain vary from about 12-26dB.

If you look back at the Tina simulation of the simple mixer shown in Fig.8.5, you will see that an IF output of around 20V peak-to-peak is being produced by an input of 1V peak-to-peak, corresponding to a conversion gain of 20 (or about 26dB).

Amplification is more readily accomplished at lower frequencies, which is why the majority of gain in a superheterodyne receiver is provided within cascaded intermediate frequency (IF) stages (following the mixer).

Gains typically range from 80-100dB. As a result, mixers are generally not required to deliver substantial amounts of gain.

Diode mixers

Simple mixers can be based on low-cost switching diodes, as shown in Fig.8.8. Although a single diode can be used, the balanced arrangement reduces the level of local oscillator injection reaching the IF output. In the arrangement shown in Fig.8.8, the RF input is at 50MHz, with the local oscillator at 59MHz.

The 9MHz output is selected by T2, which also provides impedance matching for the 9MHz filter at the output. Unlike the Mosfet mixer that we met earlier, this type of mixer requires significantly greater drive power.

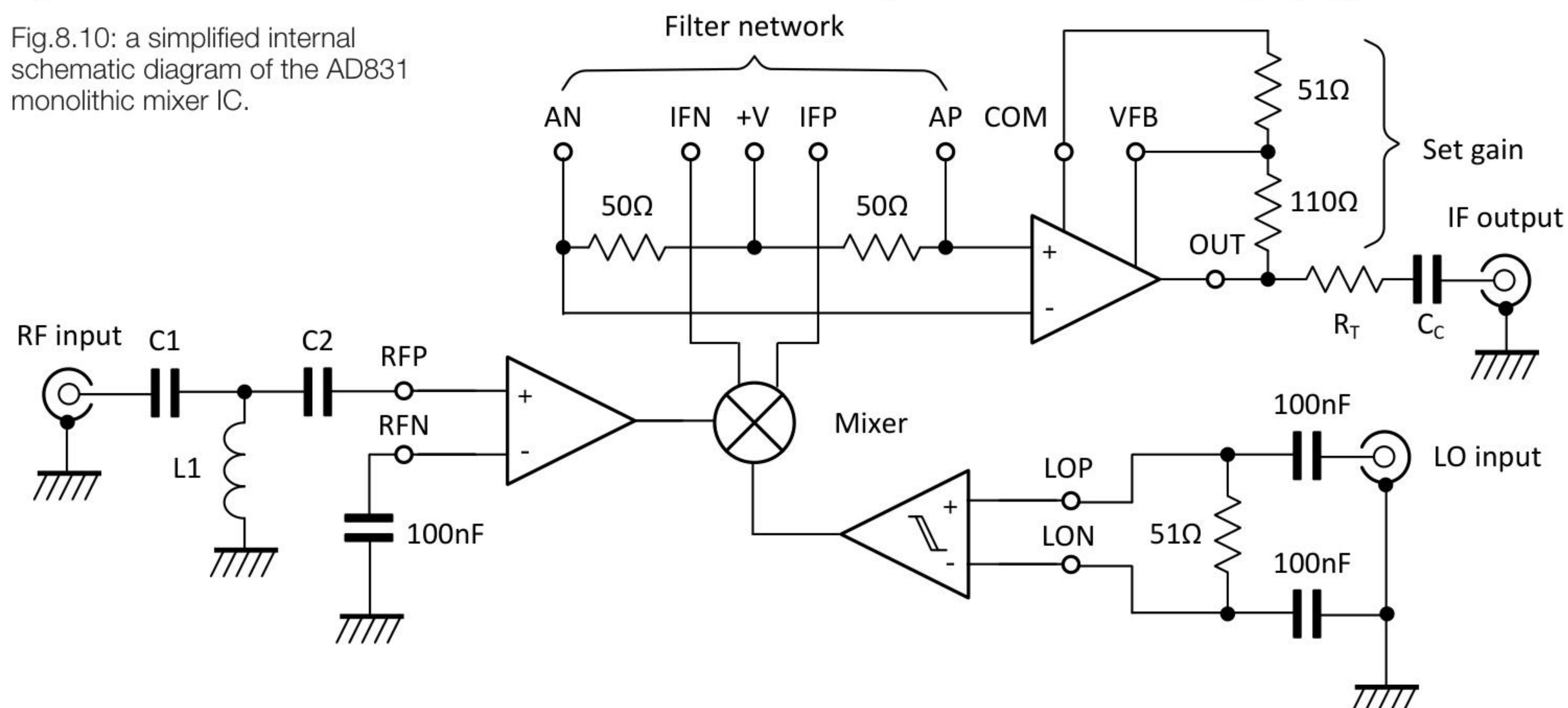
It also provides attenuation rather than the modest amount of conversion gain associated with the earlier mixers shown in Figs.8.4 and 8.6. Losses associated with diode mixers vary from about 6-9dB (ie, a 'conversion loss' rather than a 'conversion gain').

A double-balanced diode mixer is shown in Fig.8.9. This arrangement provides rejection of both the local oscillator and RF signal components present at the IF output. The two transformers are often untuned, and the broadband characteristic makes this type of mixer useful over a very wide range of frequencies.

Monolithic mixers

LSI technology is increasingly being used to create low-distortion, wide dynamic range mixers with excellent characteristics.

One such example is the AD831, shown in Fig.8.10. Typical applications include RF-to-IF down-conversion in HF and VHF receivers, second mixers in digital mobile radio (DMR) base stations, direct-to-baseband conversion, quadrature modulation, demodulation and even Doppler shift detection in ultrasound imaging applications.



The AD831 incorporates a low-noise IF output amplifier as well as a local oscillator driver. It consists of a mixer core, a limiting amplifier, a low-noise output amplifier and a bias circuit. The device has differential RF and local oscillator inputs: RFP, RFN and LOP, LON, where the Ps and Ns indicate 'phase' and 'antiphase' respectively.

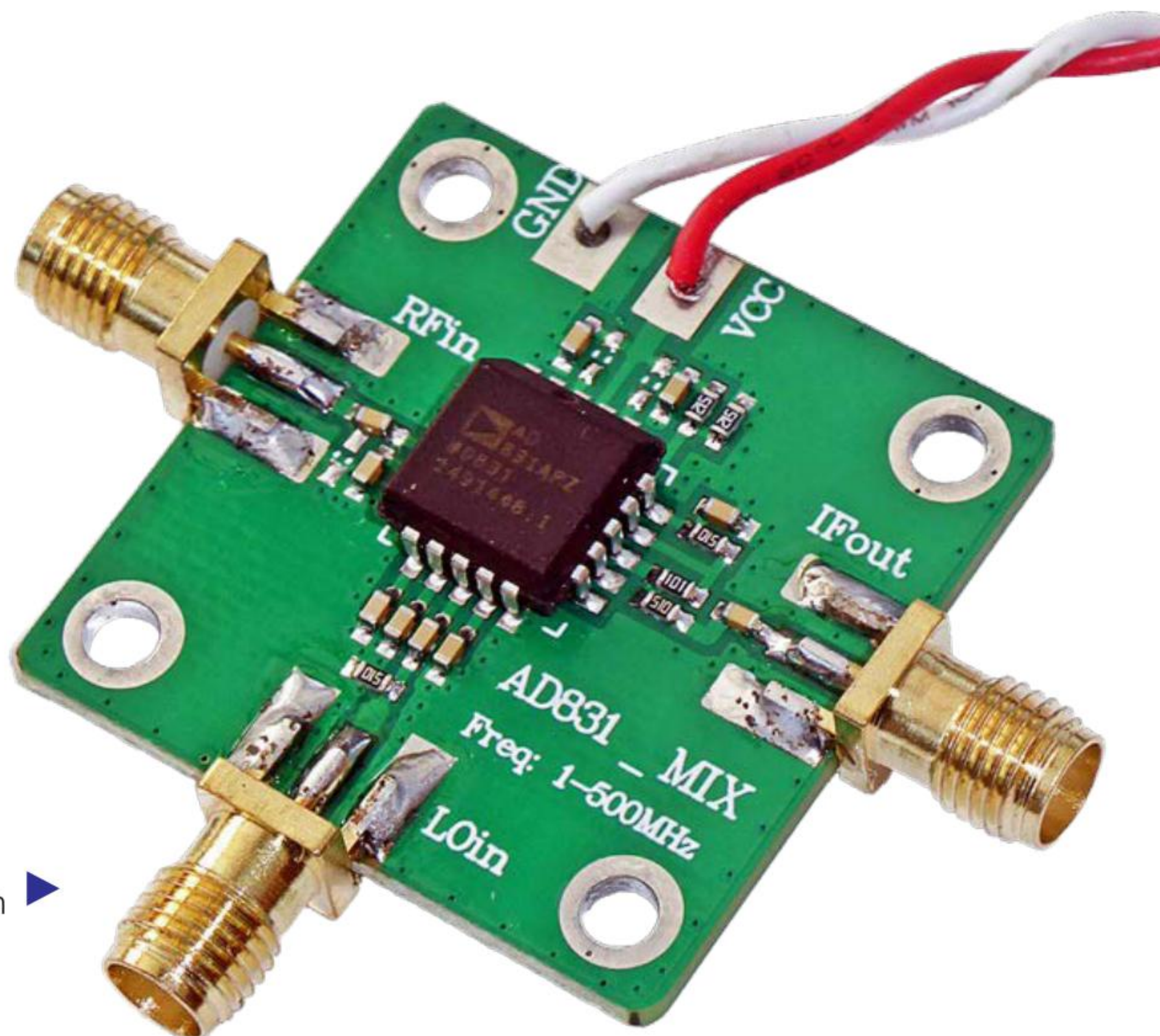
The mixer's differential outputs at the sum and difference frequencies of the RF and local oscillator inputs are available at IFP and IFN. These open-collector outputs can supply an external filter network (not shown in Fig.8.10) or can be connected to internal loads that convert the differential output currents into differential voltages that can be fed to an external IF amplifier.

When the RF and LO inputs are AC-coupled (as shown in Fig.8.10), an on-chip network supplies bias current for the RF and local oscillator stages. This network can be disabled when the mixer is DC-coupled to the external RF and local oscillator stages.

The internal low-noise amplifier stage has inputs AP and AN. When the amplifier is used, the AN and AP pins are supplied directly from pins IFN and IFP. Two on-chip 50Ω resistors are then used as loads to convert the differential mixer outputs into differential input voltages for the internal low-noise IF amplifier. The gain of this stage is set by external feedback resistors connected to the VFB pin.

The driver eliminates the need for a high-power external local

Fig.8.11: a low-cost mixer module based on the AD831 chip.



oscillator, reducing problems with isolation and shielding, and improving overall system performance. As previously mentioned, the RF, IF, and LO ports may be DC-coupled and supplied from a symmetrical $\pm 5V$ supply or, more commonly, AC-coupled and operated from a single DC supply of 9-12V.

The AD831 operates with RF and local oscillator inputs up to 500MHz and features an untuned differential output bandwidth of up to 250MHz. This makes the chip ideal for broadband applications where a large chunk of spectrum is converted from one broad frequency range to another.

A typical application might be

moving the HF spectrum into the VHF range for reception using a low-cost SDR dongle (see Teach-In Part 4 on software defined radio, February 2026). The AD831 can also serve as a second mixer in cellular and two-way radio base stations, offering better performance than conventional mixers.

Applications

To aid in construction and experimentation, the AD831 is available from several suppliers mounted on a small circuit board fitted with SMA connectors, as shown in Fig.8.11. The board is configured for AC-coupled operation and requires a DC supply of 9-12V. Fig.8.12 shows just a few of the possibilities.

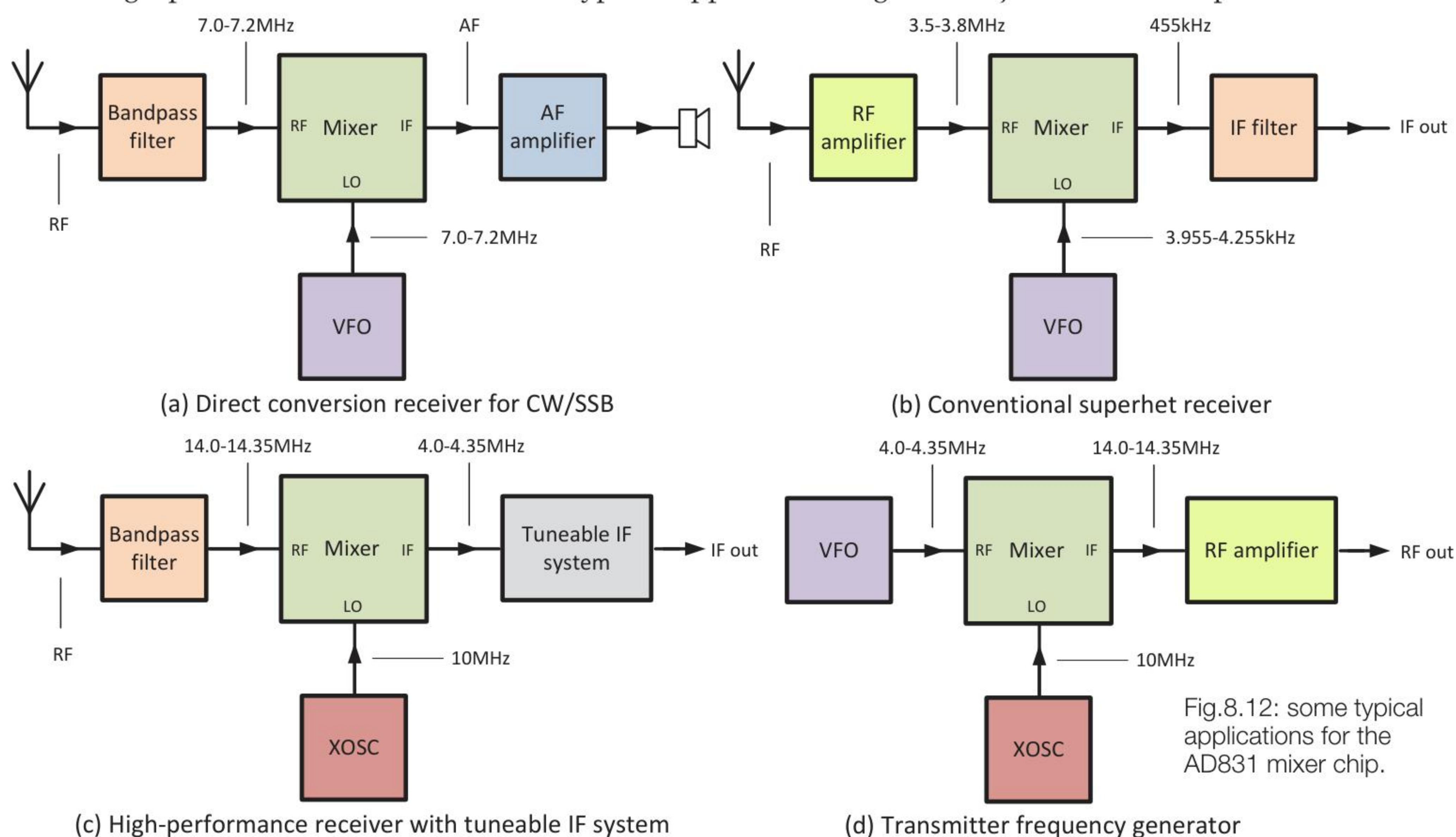


Fig.8.12: some typical applications for the AD831 mixer chip.

In Fig.8.12(a), the AD831 is used in a simple direct-conversion receiver where the local oscillator operates at the same frequency as the incoming RF signal and modulation is converted directly to baseband audio frequency. In effect, this type of receiver has zero intermediate frequency!

The bandpass filter improves receiver selectivity and reduces unwanted mixing of harmonics of the local oscillator with strong out-of-band signals.

A conventional superhet receiver is shown in Fig.8.12(b). Here, the 455kHz intermediate frequency is obtained from the difference between the incoming RF signal and the variable-frequency local oscillator. The RF amplifier provides gain and selectivity, while the IF filter reduces the level of unwanted mixing products.

Fig.8.12(c) shows an arrangement where the local oscillator is fixed and the IF amplifier is tuned over a range of frequencies. Circuits of this type are often found in dual-conversion receivers, where the first local oscillator is crystal-controlled (and therefore very stable), and a second, variable-frequency local oscillator provides tuning.

Mixers are found in transmitters as well as receivers; Fig.8.12(d) shows a typical application where the output of a low-frequency VFO is mixed with that from a crystal oscillator to produce a variable-frequency output at a higher frequency. The AD831's exceptional frequency response makes this arrangement ideal for multi-band frequency generation, where different crystals can be used for band selection.

Hands-On: A 50MHz Digital Frequency Meter

This month's *Hands-On* project is another useful item of RF test equipment, a portable 50MHz Digital Frequency Meter. The completed instrument is auto-ranging, uses a five-digit seven-segment LED display, and operates from an internal DC supply derived from four standard 1.5V alkaline cells.

To reduce cost and simplify construction, we've based the project on an inexpensive PIC-based frequency counter (see Fig.8.13). The module is available in kit form from numerous online suppliers (just search for "DIY digital frequency counter crystal tester").

Construction and operation are straightforward, but to overcome

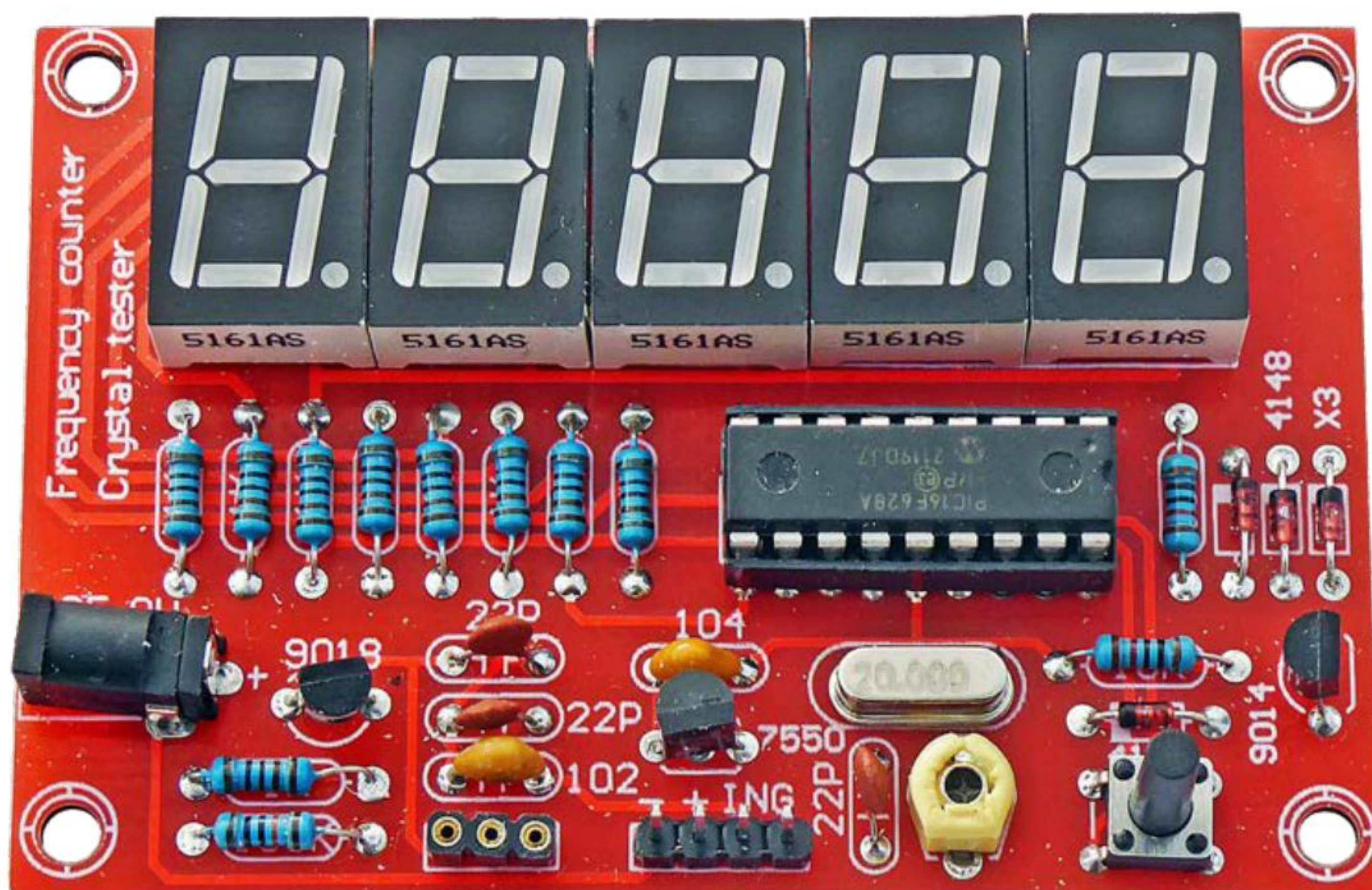


Fig.8.13: our 50MHz Digital Frequency Meter is built around this PIC-based kit.

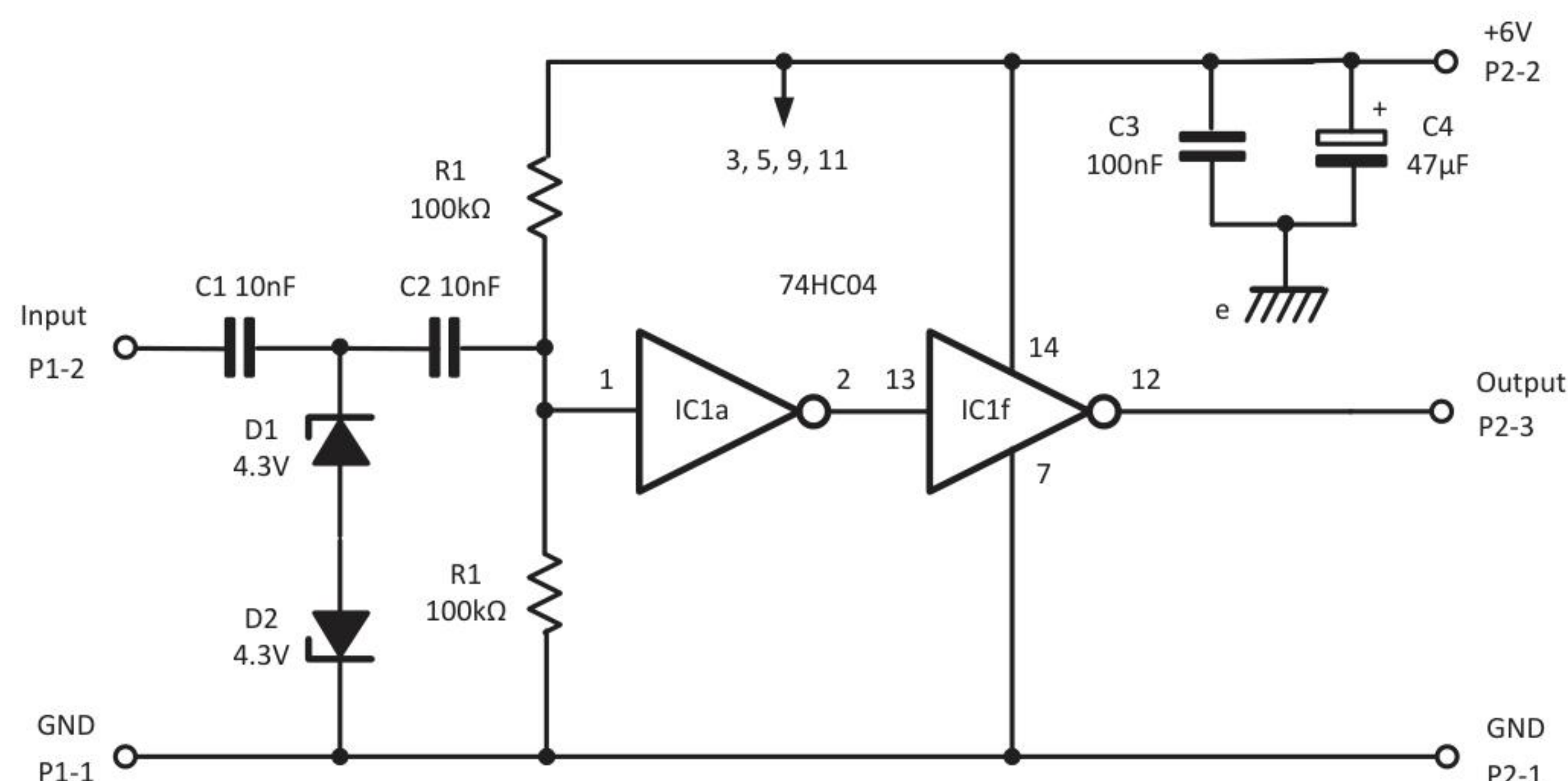
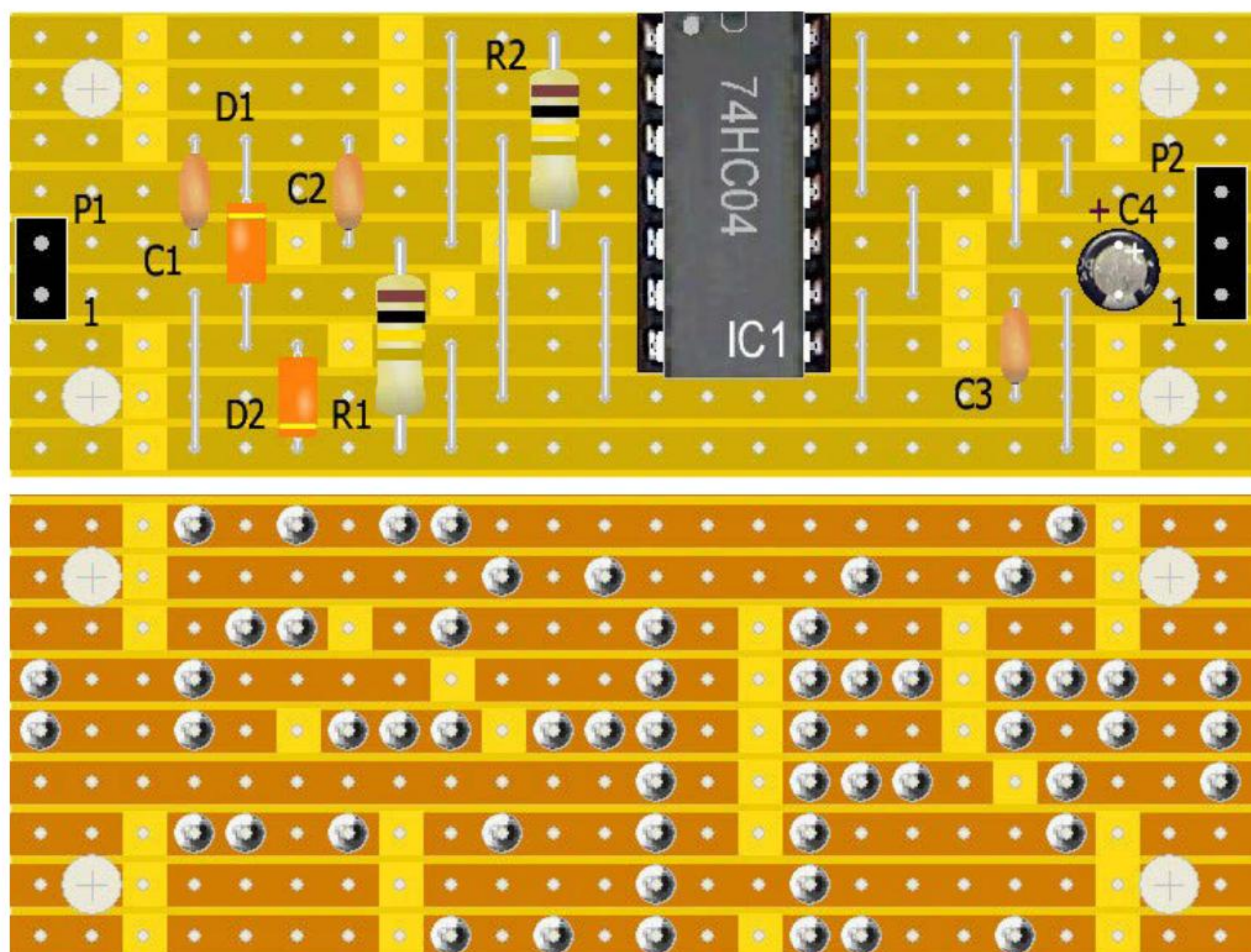


Fig.8.14: our input protection and shaping circuit for the Digital Frequency Meter.



Figs.8.15 & 16: the component and track layouts for the input protection circuit.

the lack of input protection (essential in a general-purpose test instrument), we've added a second circuit board to safeguard the in-

put and provide some additional waveform shaping.

The circuit of the input protection and shaping board is shown

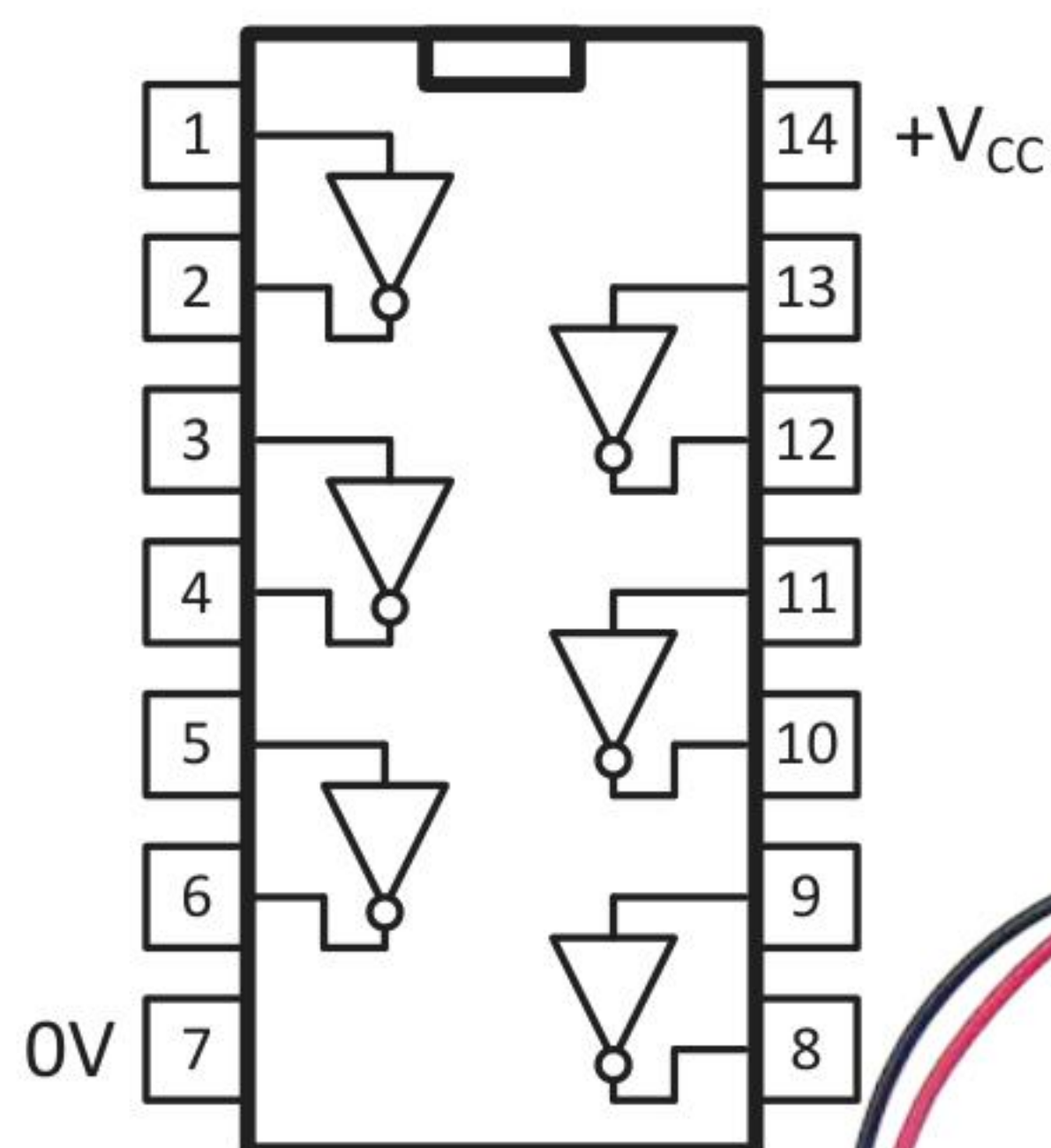


Fig.8.17: the pin connections for IC1, a 74HC04 hex inverter.

in Fig.8.14. Two zener diodes, D1 and D2, clamp the input voltage to a maximum peak-to-peak voltage of about 4.6V. The clamping arrangement is followed by a hex CMOS inverter stage with its input biased to half- supply voltage.

A sinusoidal input signal of about 150mV peak-to-peak at 1-10MHz is sufficient to produce a logic-compatible signal at the output of IC1a with fast rise and fall times. The output of IC1a is further improved by a second inverting stage, IC1f. The unused inputs of the remaining four inverting stages are all held high and their outputs left disconnected.

Assembly

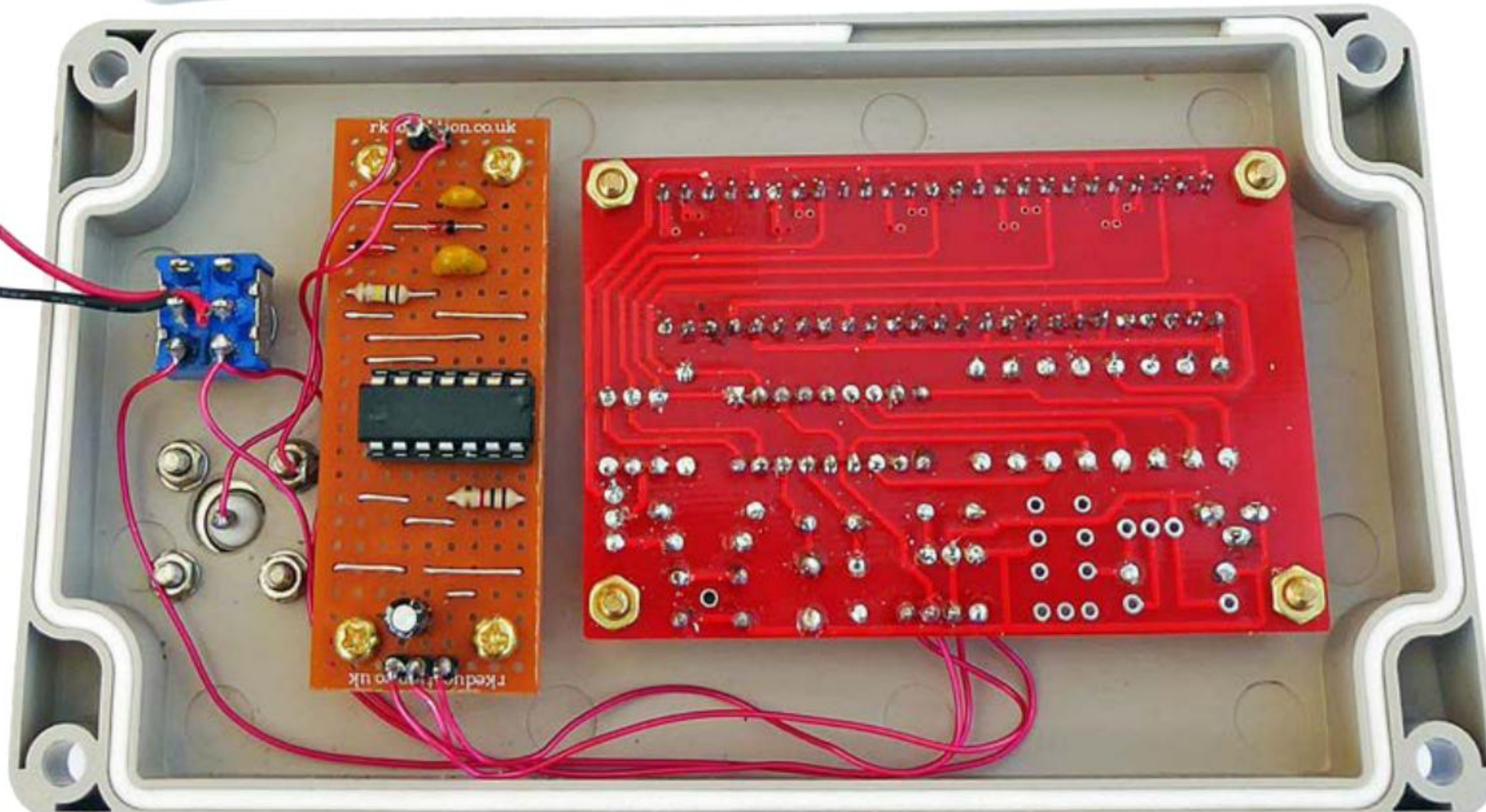
Instructions and assembly diagrams for the frequency counter and LED display board can be found on the internet at <https://www.zl2pd.com/xtalchecker.html>

The supplied PCB is clearly marked with component outlines and values, but care needs to be taken to ensure that the three transistors are identified and fitted correctly.

The external power connector (marked “DC5-9V” on the PCB) should be fitted but is not used because the internal DC supply (+6V and GND from S1a and S1b respectively) is fed directly to the “+” and “-” pins on the four-way header. The mode button supplied



Fig.8.19: wiring up the Frequency Meter.



with the kit can also be fitted but is not required for normal operation of the Digital Frequency Meter.

The component layout for the input protection and shaping circuit board is shown (viewed from the top) in Fig.8.15. The corresponding track layout (viewed from below) is illustrated in Fig.8.16. 29 track breaks are required; they can be made using a spot face cutter or small drill bit. Eleven links on the upper side of the boards can be made using short lengths of tinned copper wire.

The pin connections for IC1 are shown in Fig.8.17.

When the two circuit boards are complete, they should be mounted in the ABS enclosure using 10mm

hex spacers and interconnected as shown in Figs.8.18 & 8.19. When positioning the two circuit boards, adequate clearance must be left below them to accommodate the battery holder in the base of the enclosure. Keep the signal and ground leads for the input connector (SK1) as short as possible.

Testing and calibration

The testing procedure is straightforward, but before inserting the cells into the holder and switching it on, it's important to carry out a careful visual examination of both boards. Check the orientation of semiconductors such as diodes, transistors and integrated circuits.

When the visual check has been

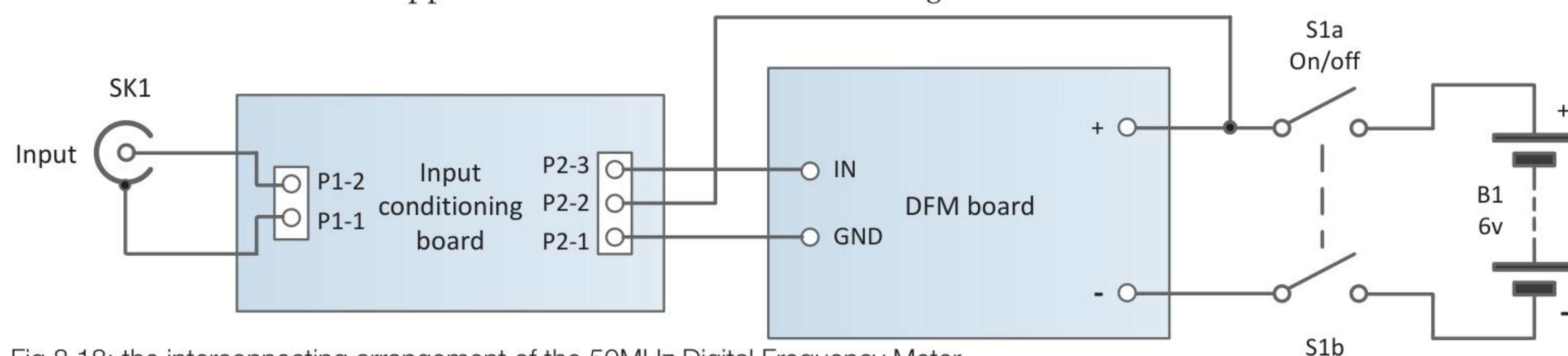


Fig.8.18: the interconnecting arrangement of the 50MHz Digital Frequency Meter.

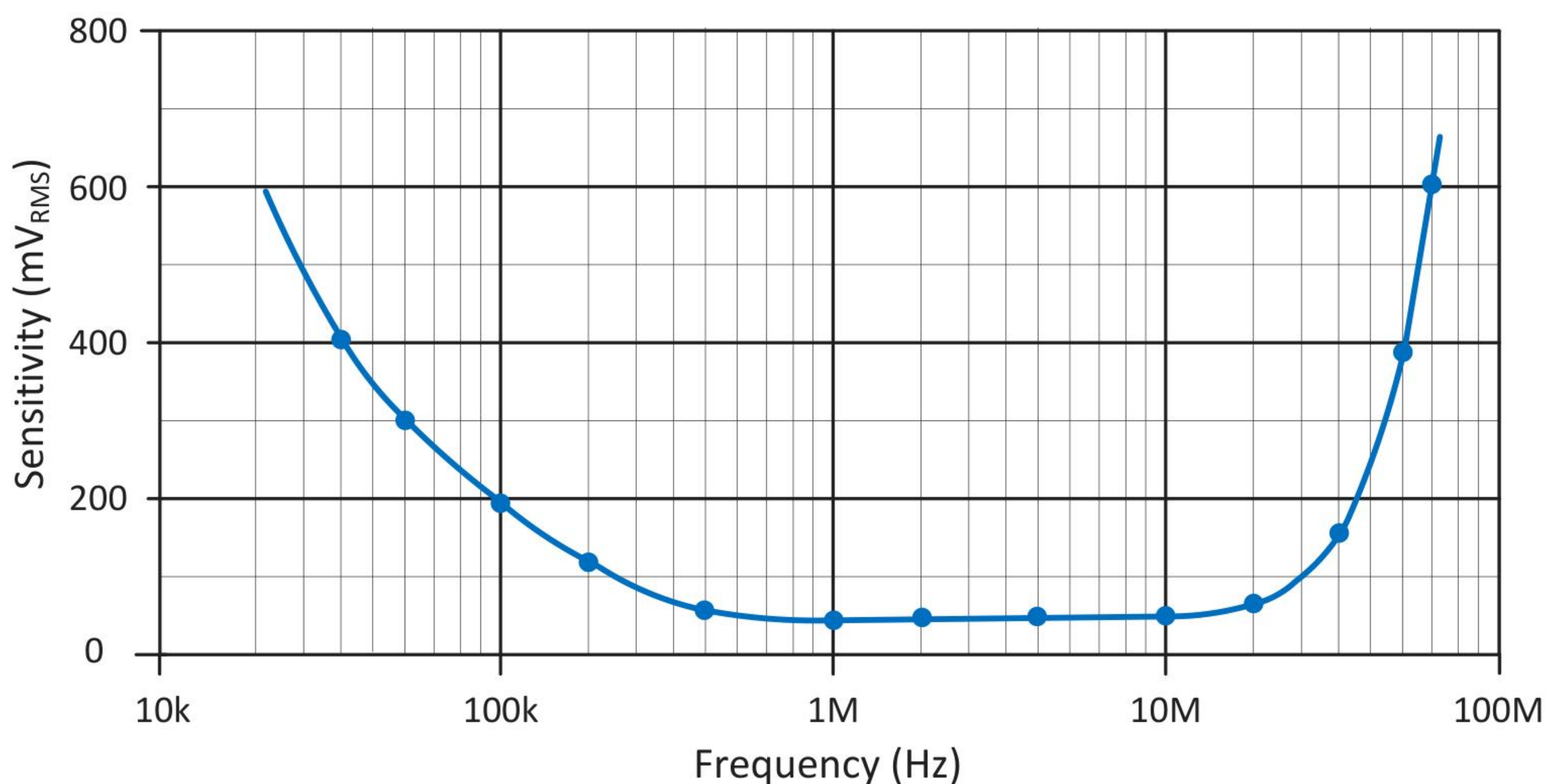


Fig.8.21: the frequency response of our prototype 50MHz Digital Frequency Meter.

completed, insert four 1.5V AA alkaline cells and switch the instrument on.

With no signal applied, a single zero should appear on the display (this reduces standby battery drain). When a signal is applied, the remaining segments will become illuminated and the supply current will increase.

Since the instrument is 'auto-ranging', the decimal point indication will change according to the input frequency range.

A small trimmer capacitor (adjacent to the 20MHz quartz crystal and 22pF capacitor) provides frequency adjustment, but calibration will not normally be necessary and should not be attempted without the aid of an accurate frequency reference.

We will be describing an accurate 1MHz/10MHz frequency reference in the next instalment of Teach-In, so if you don't have a reference source available, you might want to wait until next month before making any adjustment.

The Digital Frequency Meter is ideal for making RF measurements; the measured response of our prototype instrument is shown in Fig.8.21.

Our prototype meter produced useful indications up to around 75MHz. Its sensi-

tivity fell below 50mV RMS over a frequency range extending from 600kHz to 10MHz, and was better than 250mV RMS from 75kHz to 50MHz.

IF offsetting

The low-cost frequency counter has several features that can be accessed using the pushbutton switch, which can be made accessible by drilling a small hole in the enclosure before assembly. A momentary press of this button is needed to enter programming mode, after which you can toggle through various options by further presses.

To modify the settings, you need

to keep the button depressed for more than one second to apply the selection. The available options are "Quit", "Add", "Sub", "Zero" and "Table".

Selecting these options will allow you to display the RF input frequency of a superhet receiver while the instrument measures the local oscillator frequency. The IF offset adjustment is then made automatically by subtraction or addition. This can be a very handy feature!

To select an IF offset, you will first need to navigate to the "Table" option. Making a further button press for more than one second before releasing will allow you

Fig.8.22: the finished 50MHz Digital Frequency Meter.



to cycle through the list of IF frequencies: 455.0kHz, 4.1943MHz, 4.4336MHz, 10.700MHz and 3.9990MHz.

To select one of these, simply press the button for a further second before releasing it while the desired option is being displayed. The selected frequency can be either added to the reading (when the local oscillator is on the low side) or subtracted from it (when the local oscillator is on the high side) by returning to the main option list and selecting one of the “Add” or “Subtract” options.

Coming up!

In Part 9 next month, we take a deep dive into RF sources, explaining how to get the best out of a variety of signal and marker generator modules and direct digital synthesis (DDS) chips. Our *Hands-On* constructional project is an accurate calibration source and programmable frequency marker.

Following that, part 10 will cover RF sources and oscillators, including temperature compensation, oven-controlled oscillators and direct digital synthesis (DDS).

See you then!

PE

Parts List – 50MHz Frequency Meter

- 1 small ABS enclosure, 158 × 90 × 60mm or larger
- 1 25 × 64mm piece of stripboard (9 × 24 holes)
- 1 DIY frequency counter crystal tester kit (see text)
- 1 DPDT miniature toggle switch (S1)

Connectors

- 1 2-pin male 0.1in/2.54mm header (P1)
- 1 3-pin male 0.1in/2.54mm header (P2)
- 1 14-pin low-profile DIL IC socket
- 1 chassis-mounting female BNC socket, mounting hardware and earthing tag (SK1)
- 1 battery holder for 4 AA cells
- 1 battery clip and connecting lead

Hardware

- 8 brass or nylon M3 hexagonal 10mm tapped standoffs/spacers
- 4 M3 panhead machine screws
- 4 M3 brass hex nuts
- 8 M3 countersunk head machine screws

Semiconductors

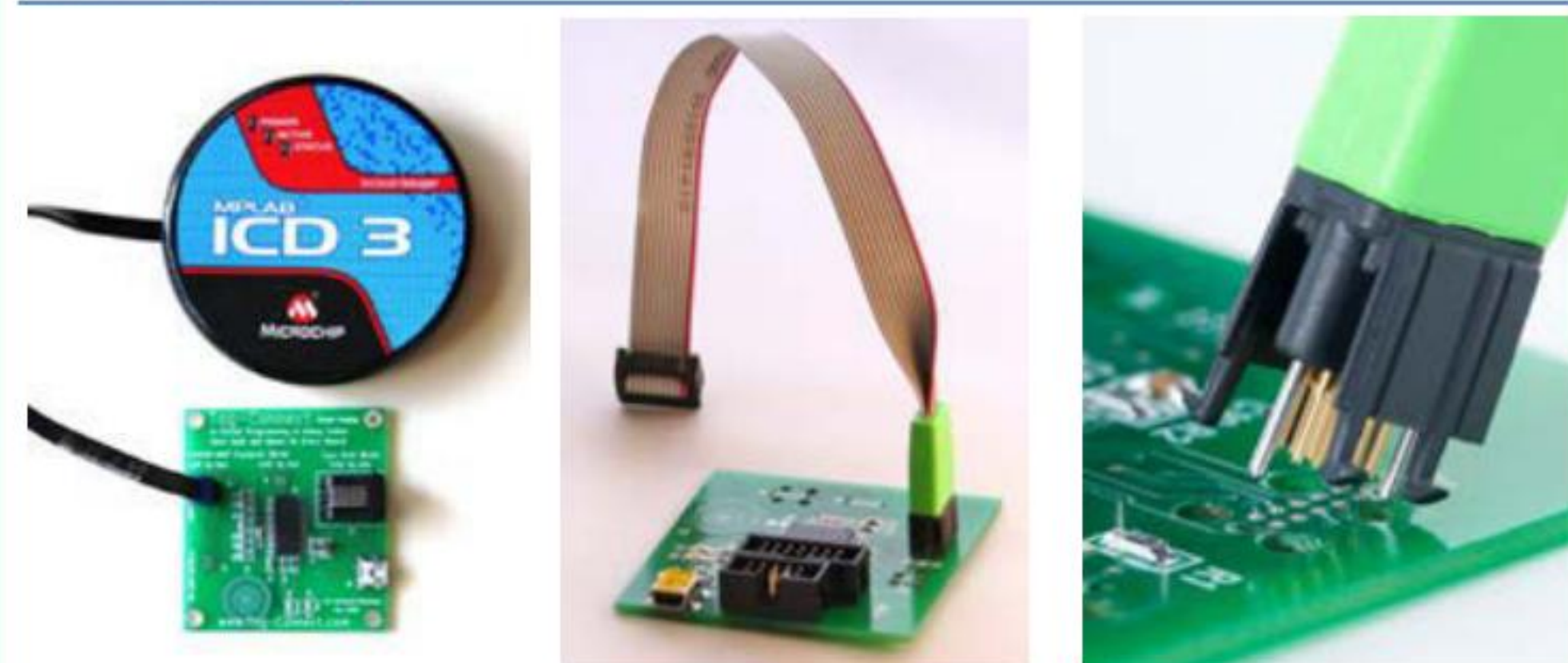
- 1 74HC04 CMOS hex inverter (IC1)
- 2 4.3V 300mW zener diodes (D1, D2)

Capacitors

- 1 47µF 16V radial electrolytic (C4)
- 1 100nF 50V ceramic (C3)
- 2 10nF 50V ceramic (C1, C2)

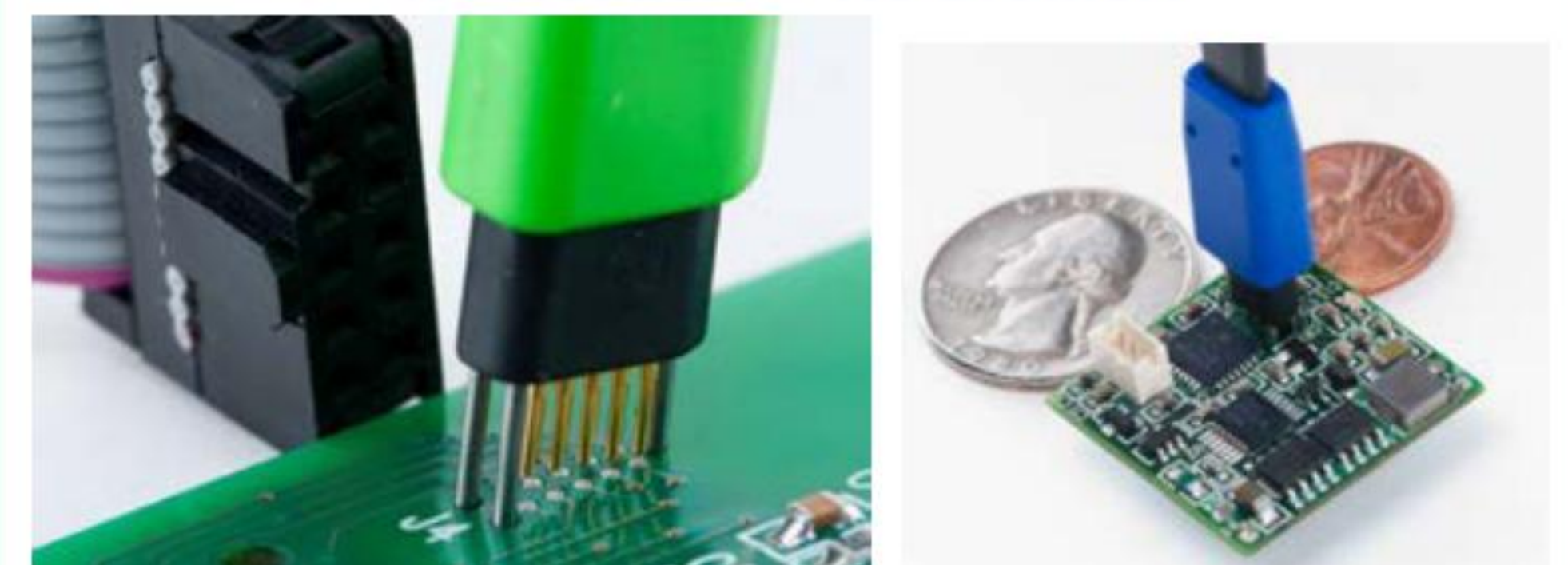
Resistors (all ¼W axial, ±5% or better)

- 2 100kΩ ¼W ±5% or better axial resistors (R1, R2)



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Constructional Project



This colourful and unique clock tells the time with LEDs arranged in a circle that light up in different colours to represent the hour, minute and second 'hands'. There's even a light chaser effect you can optionally enable that has the LEDs racing around each second. It's a straightforward build that looks great when finished.

By Nicholas Vinen

This RGB LED Clock is simple and stunning, using very few components outside the LEDs.

I came up with the idea of this clock being vaguely aware of the somewhat similar "Mesmeriser" clock by Grant-ronics that was published in the June 2005 issue of SILICON CHIP magazine. I knew it used single-colour LEDs; I thought it'd be much nicer to make my Clock from multi-coloured RGB LEDs.

This is a hybrid digital/analog clock. All the principles involved in timekeeping and the display are digital, but it mimics an analog clock in how it shows the time. Rather than two or three physical hands that rotate about the centre, pointing to the numbers, 60 LEDs are arranged around the outside of the face. They light up where an analog clock's hands would point.

A real analog clock uses different hand sizes to distinguish the hours, minute and second hands. Instead,

this clock lights up the LEDs in different colours. If they 'overlap', the colour is a mixture! For example, if the hour hand is red and the minute hand is green, when they point to the same number, the LED lights yellow (which is what you get if you combine red & green light).

Funnily enough, if you are familiar with that clock, you will see that I've had some similar ideas to its designers, the main one being that it incorporates an optional 'chaser' that runs around the clock face once per second. It's very eye-catching and also serves to indicate to the beholder the passage of each second.

Having said that it's similar to the Mesmeriser, my chaser does operate a bit differently. You can see a video of the clock in operation with this feature enabled at siliconchip.au/Videos/RGB+Clock

If you don't like or want the chaser, it's easy to turn off with a button press.

One major difference from the Mesmeriser is in terms of complexity. I wanted this clock to be as simple as possible ("keep it simple, stupid!"). So besides the LEDs, there's just one microcontroller, 20 resistors and a few other bits and pieces. That earlier design used way more parts and would have been a lot of work to build. I would say this one is elegant in its simplicity.

I also wanted to keep the parts relatively cheap so that it could be made into an affordable kit. I considered using SMD RGB LEDs, but the 5mm through-hole types I ended up using work extremely well and are quite inexpensive in bulk. The only expensive part is the PCB because it's quite large at 200mm in diameter. Still, you wouldn't want a wall clock much smaller than that.

I think the KISS (keep it simple, stupid) principle worked out quite well because this design is considerably easier to build than the almost 20-year-old Mesmeriser design. That one had a digital display as well, but I thought most people would want either a digital or analog like readout, not both. One concession I made to digital clocks is to add optional AM/

RGB LED Analog Clock Complete Kit (SC7416, about £40 + P&P)

Sold by SILICON CHIP and sent from Australia, it includes all the parts in the list except the power supply. The microcontroller is pre-programmed. Choose a BZ-121 GPS module or Pico W (unprogrammed) for the time source.

RGB LED 'Analog' Clock

ANALOG' CLOCK

- **200mm black clock face with 60 RGB LEDs that light different colours for the hour, minute and second**
- **12 different colour schemes to choose from**
- **Optional 'subsecond' hand chaser**
- **Adjustable, automatic PWM-based dimming using an LDR to sense the ambient light level**
- **Two optional single-colour LEDs (any colours) to indicate AM or PM**
- **Time source: GPS module or NTP time via the internet using WiFi**
- **Accuracy: typically within one second**
- **Time zone: from GMT-14 hours to GMT+14 hours in 15 minute increments**
- **Daylight saving: manual one hour toggle**
- **Power supply: 5-12V DC @ 50mA from plugpack or USB**
- **Time source baud rate options: 4800, 9600, 19200, 38400, 57600, 115200**



PM indicator LEDs to make the time unambiguous.

While I think this clock is nicer than the Mesmeriser, the kit is less than half the cost (even less if you consider 20 years worth of inflation!). It is also a bit larger than the Mesmeriser, and since the LEDs are right at the edge, the display is larger again.

I didn't think it was realistic to power a clock that uses LEDs for time display from a battery, so it's simply powered from a DC supply between 5V and 12V (ideally in the range of 6-9V). That could mean a plugpack or USB supply. The simplest way to connect it is to have a thin figure-8 wire hanging down from the bottom of the clock to the nearest power point.

If you really wanted to power it from a battery, with the average current draw of around 25mA for our prototype, you could expect four AAs to power it for around five days. So you'd want to use high-capacity rechargeable (eg, NiMH) cells to avoid spending a fortune on alkaline cells.

In terms of timekeeping, I have offloaded that to your choice of either a GPS module or a WiFi module that fetches NTP time via the internet and

your WiFi network. That makes the clock extremely accurate, with no drift, while also helping to keep the circuit simple.

The Clock is designed to hang on the wall as a PCB assembly. It might get a bit dusty, but you don't normally touch clocks often, so it doesn't strictly need a case. If you want to put it in a case, you can likely find a suitable one. The easiest solution is probably to buy a cheap clock in a plastic case that's large enough, gut it and install this PCB in that case. Just make sure it'll fit first!

Circuit description

The full circuit is shown in Fig.1. The key to keeping it simple is the use of Charlieplexing to allow all 60 RGB LEDs, containing a total of 180 junctions (plus the two optional AM/PM LEDs) to be driven from a 20-pin IC.

This technique involves keeping most of the pins connected to the LEDs in a high-impedance state (eg, configuring them as digital inputs). One output is driven high (to +5V) and the other is pulled low (to 0V). A different LED junction is connected across each possible pair of pins, so depending on

which pin is high and which is low, only one lights.

Due to the way the connections are arranged, we can actually light any combination of the red/green/blue junctions in a single LED at any time.

By multiplexing them (switching quickly between different states), we can make it appear that multiple LEDs are lit at once. In the case of this clock, we need to light up to five LEDs at once: for the hour 'hand', minute 'hand', second 'hand', the optional 'subsecond hand' (which goes around the face once per second) and the optional AM or PM LED.

That means each LED is lit for a maximum of 20% of the time. To compensate for that, we drive them pretty hard, so they still look quite bright. By lowering the amount of time they are lit for (ie, having all LEDs off sometimes), we can control the brightness, too.

The formula for the number of LEDs (ℓ) that can be driven by a Charlieplexed arrangement for a certain number of pins, n , is $\ell = 2(n-1) + 2(n-2) + 2(n-3) + \dots + 2$, which can be simplified to $\ell = n(n-1)$. This is a quadratic equation, so we can solve it for n and get the formula $n = (1 + \sqrt{1 + 4\ell}) \div 2$.

Plugging in our value of $\ell = 182$, we can determine we can do it with exactly 14 I/O pins. However, our circuit uses 15 because that allows us to use a much simpler arrangement where figuring out which pins to drive high and low to light any given LED is trivial. You can see the arrangement we used in the circuit diagram.

The RGB LEDs are grouped in sequential sets of four, having their common anodes all tied together and connected in turn to I/O A (LED1-LED4), B (LED5-LED9) etc, up to the 15th I/O, O (LED57-LED60).

For the first group with their anodes driven by I/O A, the cathodes are driven by B, C, D, ... K, L and M. For the second group with the anodes driven by I/O B, the cathodes are driven by C, D, E, ... L, M and N. After the third group, which ends with the final cathode being driven by the 15th I/O, O, it wraps around to A again.

The critical thing is that no I/O pin appears twice in the same group. The AM and PM LEDs are connected to spare combinations of pins that are not used for any of the RGB LEDs.

To light one LED, all we have to do is figure out which group it is in and drive the corresponding shared anode pin high. We then pull one, two or three of the cathode pins connected to that LED low to light it with a particular colour.

The nets designated A through O connect to pins on microcontroller IC1 via 68 Ω current-limiting resistors. These were calculated with the microcontroller's absolute maximum current limit per pin of $\pm 25\text{mA}$, as well as the typical limit for an LED being 20mA (although, with duty cycle always being under 50%, it isn't really a concern).

Each LED will have two resistors in series when it is lit, one in the anode circuit and one in the cathode circuit. Assuming the lowest LED forward voltage at 20mA is 1.8V and the supply is exactly 5V, that means there will be 3.2V across the resistors, allowing a maximum of 23.5mA to flow ($3.2\text{V} \div [68\Omega \times 2]$). However, the microcontroller's output transistors also have an inherent resistance that we can calculate as being close to 68 Ω from information in the data sheet.

This means that the series resistance for each LED is effectively around 200 Ω , so the actual current limit is closer to 16mA, comfortably under the 25mA limit.

Brightness adjustment

We want the LEDs in the clock to be bright in a well-lit room but not so bright at night so they don't sear your eyeballs. Thus, there is a light-dependent resistor (LDR) near the middle of the clock face that senses the ambient light level. It forms a divider with the 100k Ω resistor in series with it, across the 5V supply rail.

At higher light levels, the LDR's resistance is low, so the RA5 analog input of IC1 will be close to 5V (probably around 4.5V). As the light level drops, its resistance will increase to 100k Ω and above, so that voltage will drop to 2.5V and lower. The microcontroller can thus use its analog-to-digital converter (ADC) to read the voltage on that pin and adjust the LED brightness.

That is done using pulse-width modulation (PWM). We'll explain how it's implemented in the software section.

Cheekily, we also use pin 2 of IC1 to sense presses of pushbutton switches S1 and S2. They are used to change the clock's configuration, set the time zone, compensate for daylight saving, adjust the LDR sensitivity and so on.

When one is pressed, it pulls pin 2 either almost all the way up to 5V or all the way down to 0V. The LDR's resistance doesn't vary enough to allow the voltage to get that close to either rail, so the microcontroller knows a button has been pressed. The 220 Ω series resistors are low enough not to interfere with that, but high enough to avoid damage in case both buttons are pressed (an invalid condition).

Timekeeping

While IC1 has an internal oscillator, it isn't super accurate, so it's only used for timekeeping from second-to-second. For longer intervals, we rely on the time source: either a GNSS (eg, GPS) module, which gets its time from atomic clocks in satellites, or a module that fetches the time via internet NTP servers using a WiFi network.

Either way, we're relying on that module to have a crystal for reliable timekeeping, and we simply get its updates (once per second or more frequently) and display whatever time it gives us.

The GNSS or NTP module is connected via six-pin header CON3. Some modules have four or five wires, in which case some of these pins are not connected. Actually, only three are required: V_{CC} (5V) and GND (0V) to

supply the module with power, and TX, which is the pin it uses to send serial data to our microcontroller that includes the time.

We provide a pad for soldering the module's RX wire that pulls it high via a 10k Ω resistor so that the module doesn't get spurious serial data due to EMI. We never actually need to send it data. The 1PPS pad is provided as a place to anchor a 1PPS wire if the module has one; we don't need that either. The EN pad will pull up the module's EN wire, if it has one, to enable it.

Remaining circuitry

IC1 has a 100nF supply bypass capacitor for stability, plus a 10k Ω pull-up resistor on its MCLR (reset) pin to prevent spurious resets. Optional in-circuit serial programming (ICSP) header CON2 provides a way to re-program IC1.

While some LEDs will light dimly while doing this (as pins 18 & 19 of IC1 are used for both programming and driving LEDs), we didn't find this interfered with programming the chip.

All that remains is the simple linear power supply. REG1 is a low-dropout 5V regulator that allows you to feed in a higher voltage (6-12V DC) and it will provide a nice, stable output to run the clock. As the current draw is usually less than 50mA, it will only dissipate 350mW at most ($[12\text{V} - 5\text{V}] \times 0.05\text{A}$). It's in a medium-sized SOT-223 package soldered to the board, so it will handle that easily.

The AMS1117 regulator requires a 1 μF ceramic capacitor on its output, plus an input bypass capacitor of at least 100nF, so we have provided those. They are both 1 μF to simplify construction.

Mosfet Q1 provides reverse polarity protection, as the power input is via a simple two-pin header or soldered wires. If the supply polarity is connected correctly, Q1's gate is pulled up, and it switches on, connecting the supply negative wire to circuit ground. If the supply wires are swapped, Q1's gate is pulled down and its body diode is reverse-biased, so no current will flow.

ZD1 protects Q1's gate in case a negative voltage exceeding 16V is applied to the board, however unlikely that is.

Options

While we are specifying common-anode RGB LEDs, it will actually work

RGB LED 'Analog' Clock

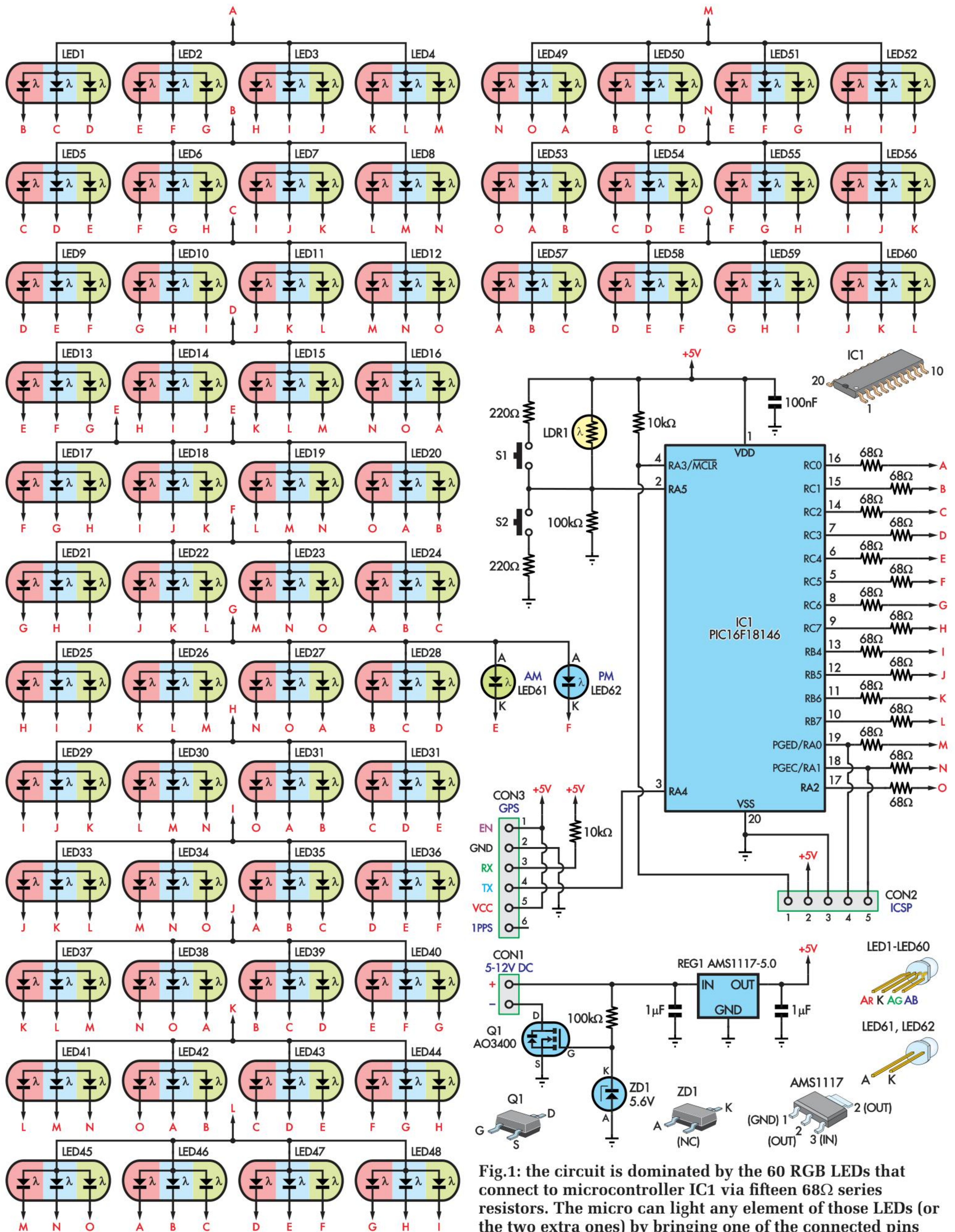


Fig.1: the circuit is dominated by the 60 RGB LEDs that connect to microcontroller IC1 via fifteen 68Ω series resistors. The micro can light any element of those LEDs (or the two extra ones) by bringing one of the connected pins high and the other low, while the remainder are kept as high-impedance digital inputs.

Constructional Project

with common-cathode RGB LEDs too. All that needs to change if CC LEDs are used is for the drive polarity to be reversed, ie, instead of pulling a pin high, it is pulled low, and vice versa.

The only catch is that if you fit the AM & PM LEDs (LED61 & LED62) and are using common-cathode RGB LEDs, they need to be reversed (rotated by 180°) so that their polarities match the other LEDs.

PCB layout

You would think with all these LED connections, the PCB layout would be a nightmare, but actually, it was straightforward. We have kept it as neat as possible, and quite symmetrical, since the PCB also forms the clock's face – see Fig.2. We could have hidden most of the circuitry on the back, but we thought it'd be more interesting to have some on the front!

The microcontroller, IC1, is right at the centre of the face, which seems appropriate since it's also logically at the centre of the circuit. Most of the resistors are to its left and right. The bypass capacitor is above it, while the LDR is centred below it. The pushbuttons are on the left and right sides, lined up with IC1.

The power supply components mount on the back of the PCB, towards the bottom.

We initially thought of using the auto-router to make the 244-odd connections to the LEDs, but came up with a better (and much neater!) idea.

The 15 I/O lines (labelled A through

O on the circuit diagram) are assigned to 15 bottom-layer circular tracks that run at fixed intervals inside the circle of RGB LEDs. Top-layer radial linear tracks run from each RGB LED pad part-way towards the centre of the PCB, all terminating just above the innermost bottom-layer ring track.

With this arrangement, all we needed to do was place one via on each of those radial tracks at the appropriate location to join it to the correct circular track (A-O). One of each of the radial tracks connected to the A-O lines is then routed to the series resistor that connects to the appropriate microcontroller pin. (Thanks to Tim Blythman for his help in suggesting this neat arrangement!)

While there are pads for the three connectors to be inserted from the front side of the board, we thought it would be a

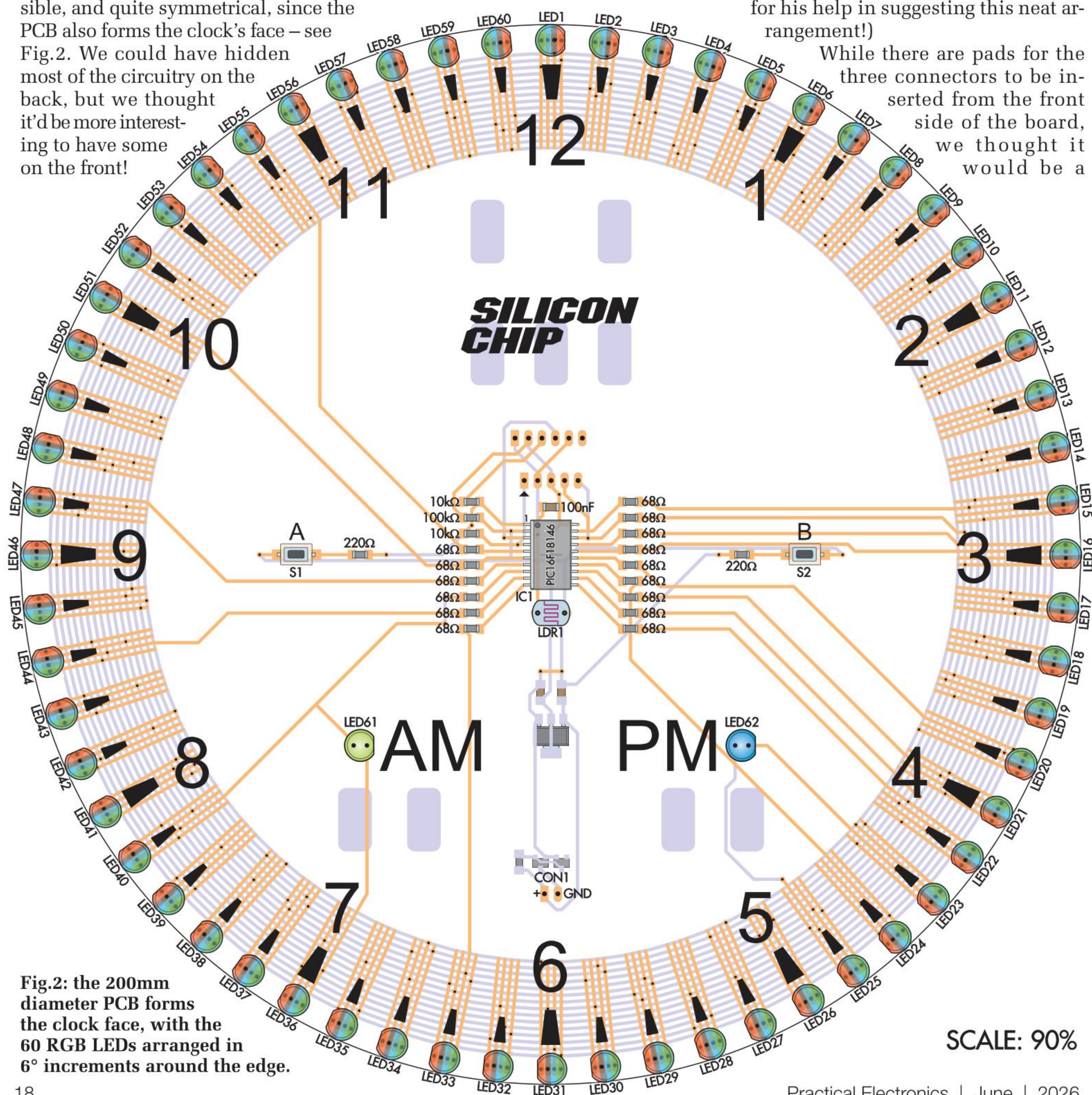


Fig.2: the 200mm diameter PCB forms the clock face, with the 60 RGB LEDs arranged in 6° increments around the edge.

SCALE: 90%

RGB LED 'Analog' Clock

bit ugly having them stick out. These pads don't have exposed copper on the front of the board, to keep it looking clean, but it is still possible to solder vertical or (ideally) right-angle headers on the back side.

Most constructors would not need to fit the ICSP header. Since GPS modules usually come with plug-in wire assemblies, there's no real need to fit a header for CON3. Instead, we suggest you simply solder the wires to those pads on the back, as we did. We stuck the GPS module on the back of the PCB with double-sided tape to hide it, as otherwise the wiring will look a little messy.

That just leaves the power input, CON1. You have a few options there. You could solder a right-angle polarised header to the back of the board and use a plug to connect it. You could also just solder bare wires (either directly from a plugpack, or to an inline barrel socket) to those pads.

There's also the option of using a small, separate board we developed that can be soldered to the main board using a pin header, which has an onboard USB socket (described starting on page 25). You can then use a 5V USB supply to power the clock. While that will probably mean REG1 will be in dropout and not regulating, and the LEDs might not be quite as bright, we haven't found it to make that much difference.

That small add-on board will be useful in many applications, so we're presenting it as a separate project.

Software

The software is simple in principle, although it is actually quite involved when examined in detail. The microcontroller runs with a 16MHz main clock ('Fosc') that results in a 4MHz instruction clock. Three hardware timers are used: TMR0, TMR1 and TMR2. The other peripherals we need are the analog-to-digital converter and the UART for serial reception.

TMR1 is used for timekeeping and to control the main loop rate. It is a 16-bit timer running from $F_{OSC} \div 4$, and it uses a 1:1 prescaler, meaning it overflows at a rate of 61.035Hz ($16\text{MHz} \div 4 \div 65536$).

Happily, this is almost exactly what we want. Assuming the subsecond hand/chaser is enabled (and it looks cool, so why wouldn't you?), the subsecond hand makes 61 steps each second. That's because it has

to go around the clock face once (60 LEDs) plus advance one step with the second hand (plus one LED). This is pure coincidence, but it works out really well!

While we advance the clock to the next second after 999.4ms ($61 \div 61.035$), we expect the GPS/NTP module to give us a time update after exactly 1000ms, so that event is used to reset the timer. That way, as long as we are getting time updates every second, there are no glitches in the time display.

TMR0 and TMR2 are used for LED multiplexing and PWM, respectively. They are both 8-bit timers that run from $F_{OSC} \div 4$, using a 4:1 prescaler, so they run at 3.906kHz ($16\text{MHz} \div 4 \div 4 \div 256$). Each time TMR0 overflows, it triggers an interrupt routine that switches to lighting the next LED (using port masks precalculated by the main loop).

That means each LED is lit for 51µs at a time, giving a duty cycle of approximately 20% at full brightness.

TMR0 and TMR2 run in lockstep. At full brightness, TMR2 never triggers. The PR2 register controls how soon the TMR2 interrupt occurs after TMR0 overflows. As the value in PR2 is reduced, the TMR2 interrupt occurs earlier and when its interrupt service routine (ISR) is called, it simply switches all LEDs off. The earlier that happens after switching on, the dimmer the display becomes.

This means we can use the PR2 register as an 8-bit PWM control for all LEDs, since it occurs at the same interval after each multiplexing step.

Besides calculating which LEDs to light based on the current mode and time, the main loop also receives serial data from the GPS/NTP module, decodes it and uses it to update the current time. After getting the UTC/GMT time, the configured timezone offset and DST offset (if enabled) is applied before updating the clock face.

It reads the voltage from the LDR divider on each run through of the loop, filters it, and uses it to calculate the new brightness level.

It also checks if button S1 or S2 has been pressed, performs debouncing, looks for long or short presses and changes the mode as appropriate.

Construction

The RGB LED Clock is built on a large circular PCB measuring 200×200mm (ie, 200mm in diameter) with a black solder mask, coded

19101251. We recommend you start by fitting the top-side SMDs, with IC1 being the first.

You should have a temperature-controlled soldering iron, fine-tipped tweezers, solder wire, a syringe of flux paste and a roll of solder-wicking braid on hand. Work in a well-ventilated area, either using a fan, next to an open window or outdoors as flux smoke is not good for you if you inhale it.

If you bought a kit or programmed microcontroller from the SILICON CHIP Online Shop, your chip will already be programmed and you can just solder it to the board. If you have a blank chip and the ability to program it off-board, do so now. We described how to do that in our article on the PIC Programming Adaptor from the September 2024 issue.

If you don't have such a programming rig and your chip is blank, solder it now and you can use CON2 to program it later.

Before soldering it, make sure you identify pin 1 and line it up with the markings on the board and in the top-side component overlay diagram, Fig.2. Fixing a reversed SMD chip is not fun! Tack one pin to the board using a little solder, then inspect the leads to verify they are all over their pads. If not, remelt that solder and gently nudge the chip into position.

Verify again that the IC is orientated correctly, then solder the diagonally opposite pin. Apply flux paste down the pins on both sides, then solder all the remaining pins. You can do it individually, by adding a little solder to the iron tip and touching it to the junction of the pin and pad, or you can drag-solder them several at a time.

If any bridges form (which is likely with drag soldering), add a little flux paste and then use solder-wicking braid to remove the excess solder.

Inspect all the solder joints and pins under magnification to ensure all pins have been soldered to their pads with good fillets and no bridge remain, then move onto the passives, using Fig.2 as a guide. It's best to fit the 220Ω, 10kΩ and 100kΩ resistors first, since all the remaining resistors will be the same value (68Ω). The sole capacitor on this side is the 100nF type.

You can use a similar technique as for the IC: tack the part to one pad, check its orientation, adjust if necessary, then solder the other pad. Add a

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tiny bit of flux paste to the first one and heat it with the iron to reflow the joint.

With all these parts in place, clean the flux residue off the board using alcohol or a flux solvent. It's much easier to clean the board before any through-hole parts are in place. This will be the clock face, so you want it to be nice and presentable! You'll probably have to clean it two or three times, using a fresh section of lint-free cloth each time, to get it looking nice.

Now solder the two tactile switches using the same technique. This will probably leave a little flux residue, but we don't want to soak the switches in solvent as it can damage them,

so apply a little solvent to a section of the lint-free cloth and wipe off the flux residue using that.

Flip the board over (rest it on a cloth so its face doesn't get dirty or scratched), then solder the parts on the reverse side, as shown in Fig.3. Don't get Q1 (possibly coded XORB) and ZD1 (likely coded T12 or Z3) mixed up, as they are in the same type of package. When finished, clean up the flux residue, although thoroughness is less crucial since it will be against the wall later.

You can now solder the LDR on the front side; you may need to bend its leads in to make them fit the pad pat-

tern. Make sure it's straight and close to the board so that it looks nice.

Soldering the LEDs

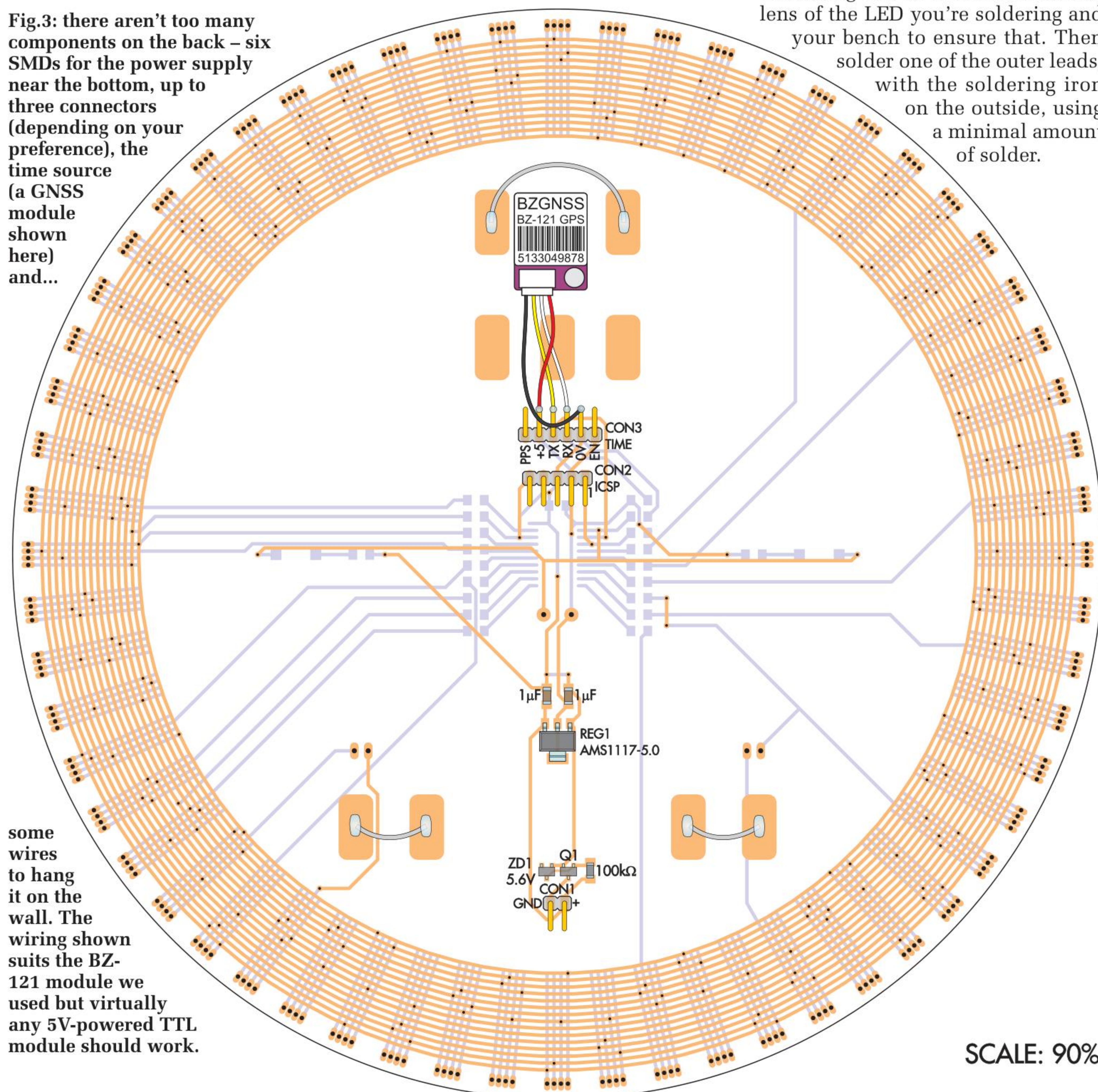
Now for the big job. This will probably take you at least a couple of hours. Don't rush it as it's harder to fix problems like bridged pads than it is to do them right the first time.

Make sure you fit each of the 60 RGB LEDs the right way around. Pay attention to the orientations of the flat faces on the PCB silkscreen and in Fig.2.

For each LED, insert it and then rest the board on your work surface such that the weight is on that LED, so it's pushed into the board. You could place something like an eraser between the lens of the LED you're soldering and your bench to ensure that. Then solder one of the outer leads, with the soldering iron on the outside, using a minimal amount of solder.

Fig.3: there aren't too many components on the back – six SMDs for the power supply near the bottom, up to three connectors (depending on your preference), the time source (a GNSS module shown here) and...

some wires to hang it on the wall. The wiring shown suits the BZ-121 module we used but virtually any 5V-powered TTL module should work.



SCALE: 90%

RGB LED 'Analog' Clock

Don't use too little; you need to be able to see that a fillet has formed. Still, you should use just enough solder to get that fillet. Next, check if the LED is flat and straight. If not, you still have a chance to push it into the PCB with your finger while you remelt that initial solder joint.

Once you're satisfied, solder the opposite lead, again with the iron coming in from the outside. The critical part here is that the pads are very close together, with large holes and thick leads, so it's quite easy to bridge the pads and somewhat difficult to fix it if a bridge forms.

For the two remaining leads, bring in the soldering iron from the middle or outside of the board so it touches the pad and lead, then feed in a minimal amount of solder from the outside until it melts, as shown in Fig.4. Again, use just enough to form a fillet. Repeat for the final lead/pad, then use sharp side-cutters to trim all four leads reasonably close to the PCB and evenly.

When trimming, use eye protection and/or hold onto the leads so they don't fly into your eye! They can attain quite a high velocity when cut through.

Now check to make sure the pads have not been bridged. If you have followed our instructions carefully, they should not be. If they are, add some flux paste and use solder-wicking braid to remove the excess solder until the bridge is clear. You can then add a tiny bit of solder back to the pads to ensure there's enough.

Repeat until you have soldered 5-10 of the LEDs.

Testing the LEDs

It's best to test the LEDs as you go, because a short circuit at any point on the board can cause the whole display to go haywire (after all, there are only 15 separate tracks connecting to all 60 LEDs). When first powered up, the firmware runs a display test where LED1-LED62 are lit up white in sequence at about 4Hz. This should allow you to quickly spot problems.

I found the easiest way to power the board was to get a regulated 5V supply (eg, a bench supply) and use clip leads to attach two male-male jumper wires to its outputs. With the clock face towards you, right-side up, rest the black (negative) wire in the right-most terminal of CON1 (the one labelled GND in Fig.2). The weight of

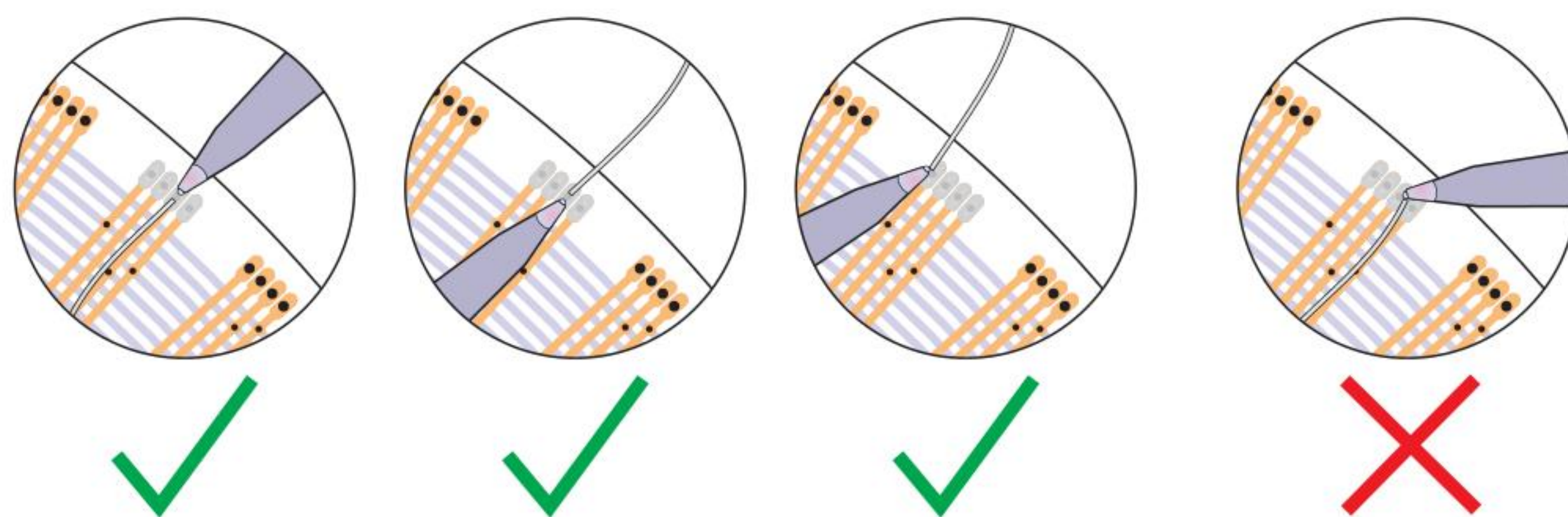


Fig.4: no doubt there are many ways to solder the RGB LEDs without accidentally bridging the very closely-spaced pads, but this is what we found worked best. The soldering iron tip and solder should come in from 180° opposite positions along the long axis of the pads. That minimises the chances of accidental bridges, which are difficult to clear. The final PCB will feature more widely spaced pads to make this a little less tricky than our prototype (although we managed to do it).

the cable will bring it contact with the plated through-hole.

Similarly, insert the positive jumper wire into the side of CON1 labelled + and ensure the two wires are not shorted, then switch the supply on. Ideally, it should be current-limited to around 50mA (the board will draw less than 20mA in this configuration). Power it up and check that LED1 lights, then LED2 etc.

If you don't have a current-limited supply, you could use a fixed-voltage DC supply with a series resistor to prevent damage in case there is a fault. For example, a 12V supply with a 150Ω 1W series resistor would easily deliver the ~20mA needed to power the circuit for testing but would be limited to 80mA in case of a fault (with the resistor dissipating just under 1W).

When powered, if any of the LEDs don't light up white, or they light at the wrong time, switch it off and check the most recent LEDs you've soldered for bridges or bad joints. Once they all light up OK, go back and solder a few more LEDs. Keep testing until all 60 are fitted and they are lighting up nicely.

You can then solder LED61 and LED62 if you are going to use them. Remember to reverse them if you used common-cathode RGB LEDs. They will light up individually during the test, after all 60 of the RGB LEDs.

While the LED solder joints are on the back of the board, if you have a lot of flux residue left behind, you may want to clean it up to avoid getting sticky hands when handling the clock. A proper flux remover is better than alcohol here, as alcohol dries fast, but you can use it if that's all you have. Because of the sharp leads, you will have to dab it off, rather than wipe it off.

It took a little while, but we cleaned this flux residue up. Try not to let any get onto the face (ie, the opposite side of the PCB), or you'll have to clean it again so it looks nice.

Final assembly

There isn't much left to complete the clock. If you're using CON2, fit it now; most constructors will not need it. We suggest you stick the GNSS or NTP module to the back of the clock using double-sided foam-cored tape. Cut it to size and stick it on. The GPS module will usually have a flat, non-conductive ceramic antenna that you can stick somewhere without too many tracks.

The NTP module has conductive parts on the back, so make sure none of them can short to the PCB. Most of the back of the Clock PCB is covered with a solder mask, so short-circuits are unlikely, but you should still check.

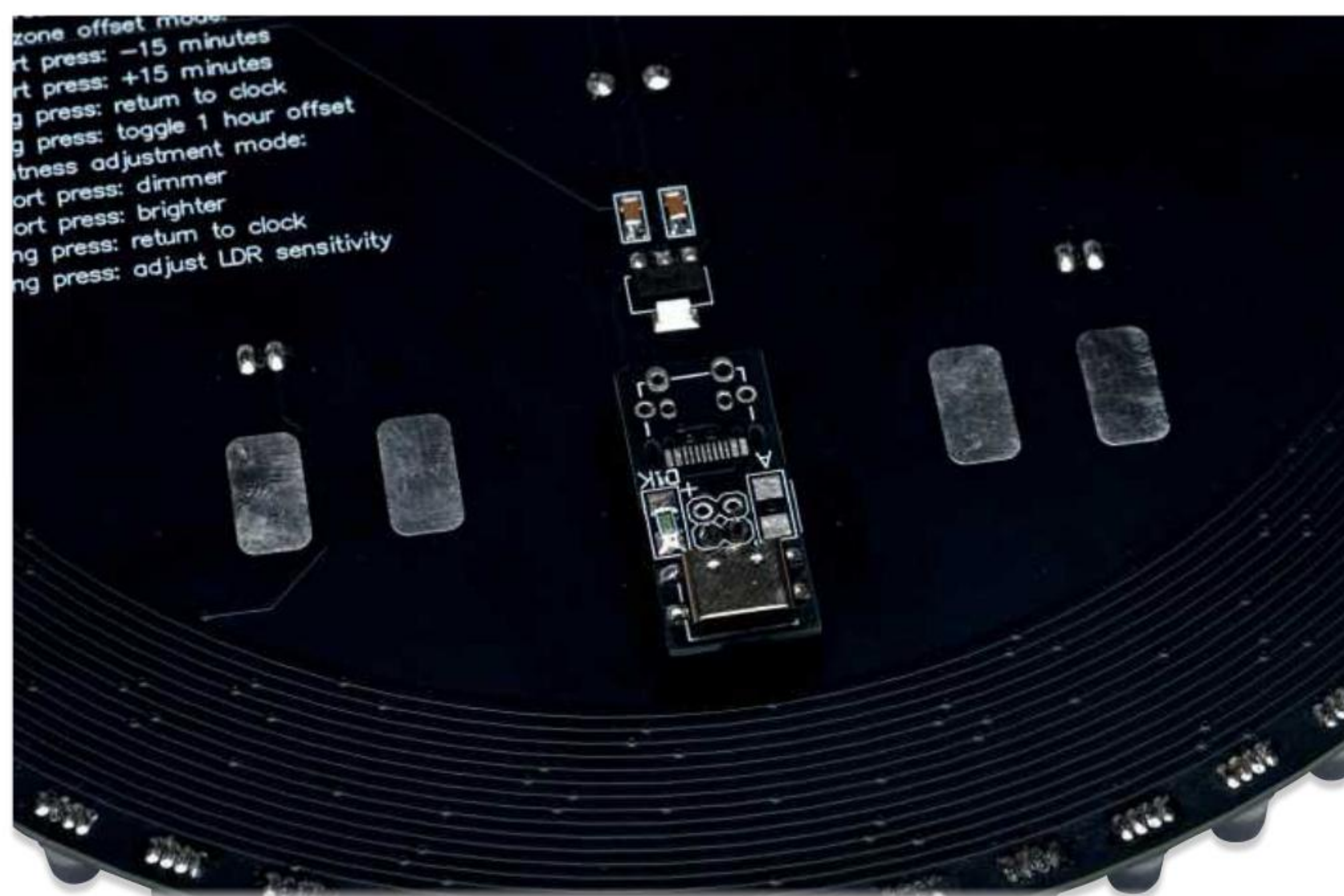
If you're using an NTP module like the WiFi Time Source for GPS Clocks that was published in the June 2024 issue, you will need to configure it so it can connect to your WiFi network. That is usually best done before wiring it up; refer to that article for instructions.

Now wire the GPS or NTP module to CON3. You can use a header and plug(s), if you want, but we think soldering wires directly to the pads is fine.

As mentioned earlier, you won't necessarily have all six wires to connect to CON3. It doesn't matter; the vital ones are VCC, GND and the TX signal from the module (which goes to the pad marked TX, not RX; ie, the labels are from the module's point of view).

If your module has an EN wire, check that it's active high. If so, you can solder

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There's nowhere suitable on the PCB to mount a USB connector, so if you want to use USB power, you can attach the add-on board like this.

it to the EN pad. Otherwise, solder it to GND. Most modules will have an RX pin, which should be soldered to the RX pad, pulling it high to disable it.

Power supply

For power, you can solder a light-duty figure-8 lead to the pads for CON1 on the rear of the board and either hard-wire it to a 6-12V DC plugpack or similar, or have an inline barrel socket or other connector at the end of that wire to connect a DC supply. Try to get the polarity correct (refer to Figs.2 & 3), although the board should not be damaged if you do accidentally reverse it.

Another option is to build our USB Power Breakout Board, described in the accompanying short article starting on page 25. This is a small PCB that can accept a USB Type-C or USB Type-B Micro/Mini socket and a pin header. It supplies ~5V DC to that pin header when a USB supply is connected. This may not run the clock at full brightness, but it won't be far off, and USB supplies are common and convenient.

We didn't put pads for a USB socket on the clock PCB itself because there was no room to do it along the edge (the whole edge is filled with LEDs). If the socket wasn't near the edge, there wouldn't be room for a cable to plug in, so the PCB would need an ugly cut-out.

Spacing this small PCB off the back of the clock PCB on a header provides enough room for the cable to plug in. The only components you need on that PCB are the USB socket, a 0Ω

SMD M3216/1206 resistor and the pin header. Once that's assembled, you can then solder it to the back of the Clock PCB as shown in the photo above.

Note how the USB Power Breakout Board PCB has four through-hole pads; this allows a two-way pin header to be fitted with multiple different polarities. The header position shown in Photo 1 is required to match the polarity of CON1. If you're unsure, refer to the separate article and its PCB overlays to see how the header position affects the polarity.

Finally, you will have noticed several large rectangular pads on the back of the Clock PCB. These are provided to solder a loop of wire for hanging the clock, with other pads near the bottom to solder wire loops so it hangs vertically on the wall.

The photo overleaf shows how we attached the wires to our Clock for hanging; you can use a similar arrangement. You can bend the wires into different positions to suit your hanger (whether it's a nail, screw, hook or whatever). If possible, do that before soldering the wires to avoid stress on the soldered joints.

Final testing

Presumably if you've gotten to this point, the LED testing went well, so the microcontroller and LEDs are working. There isn't much else to go wrong, apart from the GPS/NTP module wiring and the power supply itself.

We suggest you use the same procedure you used for testing the LEDs and

check that the current draw is under 100mA. With the BZ-121 module we used, our Clock drew around 45mA. If that's the case, it's unlikely you have anything really wrong.

You can test the LDR-based dimming now. Place a small opaque object like a credit card over the LDR and observe the LEDs. They should dim when you block the light to the LDR and brighten again when you remove the obstruction. You will have to cover the LDR thoroughly, as even a bit of light leading around its edges is enough to make the LEDs quite bright.

If that doesn't work, you probably have a soldering problem with the LDR or its series resistor, a button is stuck down or the LDR is the wrong type.

If you're using a GPS module, we suggest you put the clock near a window and set the baud rate to the correct one for your module, as described below. Leave it powered for around 30 minutes, then come back and check if it's showing the correct time for London (ie, GMT). If so, that means it has acquired the signal and is decoding the data properly. You can then complete testing by setting it up.

If using an NTP module, you should have set it up earlier, so once you have set the correct baud rate (as explained below), it should connect and show you UK time within a few seconds. In that case, proceed to the following section.

Setting it up

The first setting is the baud rate. This can only be changed after power-up when the initial LED test has completed and the spinning chaser (at roughly one 'rotation' per second) is operating, to indicate it's waiting for GPS/NTP data. If you see a clock face instead, it's likely the initial baud rate was already correct and it's getting data, so you can skip this bit.

The chaser will initially be red if it isn't getting any valid data, changing to green if there is data but no valid time yet. Once it's green, it's usually just a matter of time before it switches to telling the time (assuming you have a strong enough GPS or WiFi signal).

During this time, one of the digits 1-6 will be lit blue, indicating the baud rate:

1. 4800 baud
2. 9600 baud
3. 19,200 baud
4. 38,400 baud
5. 57,600 baud
6. 115,200 baud (default)

RGB LED 'Analog' Clock

Pressing A will go to the next lower baud rate, while pressing B will go to the next higher one. If you don't know the correct baud rate, try each one for a few seconds until the chaser changes to green. If it doesn't for any baud rate, switch off and check your wiring. Remember that TX from the module should go to the TX pad on the PCB.

All settings, including the baud rate, are stored in EEPROM, so you shouldn't have to do this again.

The remaining settings that can be accessed in time display mode are:

1. the time zone offset and optional DST (+ 1 hour when activated)
2. the colour scheme
3. the second and sub-second hand modes
4. the LED dimming calibration (minimum & maximum brightness)

Once you're in clock mode and the time has been acquired, you can set the time zone offset. Hold down the A button for a second and release it. The time display should remain, but it will now flash at 1Hz with a 50% duty cycle.

If daylight saving is active in your area, hold down B for one second to enable DST mode. A short press of the A button will make the time 15 minutes earlier, while the B button will make it 15 minutes later. Use this to set the correct local time, then hold down the A button to return to the normal display. The time zone you set will be stored in EEPROM.

After this, if your area has DST and the time changes, you just need to go into this mode and hold down B for one second, then hold down A for one second to switch between DST and non-DST. Alternatively, you could just change the time zone offset until the time is correct.

Colour scheme

In time display mode, short presses of the A button will cycle through the six possible colour schemes for the hour, minute and second hands. Each can be red, green or blue, but they must all be different. Use whatever scheme is easiest for you to remember. The default is blue for the hour 'hand', green for minutes and red for seconds.

A short press of the B button will cycle through the six possible second-hand modes. The default is to have the second 'hand' visible, plus the 'sub-second hand' in the same colour.

The sub-second hand is a chaser that runs around the clock face each second. It starts where the second hand is, goes all the way around, and 'pushes' it over to the next second on the tick.

A short press of B will cycle through the four possible second-hand modes:

1. the second hand and sub-second hand are visible and matching colours (the default)
2. the same as #1 but with a dimmer sub-second hand
3. the second hand and sub-second hand are visible, with the sub-second hand being white
4. the same as #3 but with a dimmer sub-second hand
5. the second hand is visible but the sub-second hand is not
6. there is no second hand, only the hour and minute hands

Dimming adjustments

A long press on the B button in time display mode will switch to the brightness/dimming adjustment mode. In this mode, you can control both the maximum brightness and how the brightness reduces at lower ambient light levels. By default, the maximum brightness is 100%, reducing to a low

level, but not the lowest possible, in total darkness.

Upon entering this mode, you are adjusting the maximum possible brightness. Pressing A will reduce the maximum brightness and you will see the display dim. Pressing B will increase it (if it is not at its maximum).

While making this adjustment, the LDR reading is ignored; you are seeing the brightness level that will be used at the highest possible ambient light level.

Keep in mind that, depending on where your clock is positioned, it may not normally 'see' a very high ambient light level. For this reason, you can actually set the maximum brightness above 100%. This will not make the display brighter, but it will mean that the ambient light level has to drop further before the brightness starts reducing.

The clock face shows a continuous chaser in this mode to help you see the brightness level you've set, which spans a portion of the clock face related to the possible brightness range. The portion from six o'clock to twelve o'clock shows the maximum brightness setting, so the chaser gets

Parts List – RGB LED Analog Clock

- 1 black double-sided PCB coded 19101251, 200×200mm
- 1 5-12V DC 100mA power supply (6-9V DC recommended)
- 1 2-pin vertical or right-angle header (CON1; for power – see text)
- 1 5-pin right-angle header (CON2; **optional** – for ICSP)
- 1 6-pin right-angle header (CON3; **optional** – for GPS module)
- 1 5V GPS module or compatible NTP time source (MOD1)
- [BZ-121 GPS module recommended, SILICON CHIP SC7414]**
- 2 2-pin SMD black tactile pushbutton switches (S1, S2)
- 1 20×20mm piece of foam-cored double-sided tape (to affix GPS module)
- 1 200mm length of tinned copper wire (to make hanger/standoffs)
- 1 USB power supply module (**optional**; see text and accompanying article)

Semiconductors

- 1 PIC16F18146-I/SO micro with 1910125A.HEX, wide SOIC-20 (IC1)
- 1 AMS1117-5.0 15V input low-dropout linear regulator, SOT-223 (REG1)
- 1 AO3400 or equivalent logic-level N-channel Mosfet, SOT-23 (Q1)
- 1 BZX84B5V6 or BZX84C5V6 5.6V zener diode, SOT-23 (ZD1)
- 60 frosted-lens 5mm RGB LEDs (LED1-LED60) *
- 2 5mm high-brightness LEDs with diffused lens of various colours (LED61 & LED62; **optional** AM & PM indicators)

* the kit will come with common anode LEDs but common cathode types can also be used

Capacitors (all SMD M3216/1206 size 50V X7R)

- 2 1µF 1 100nF

Resistors (all SMD M3216/1206 size unless noted)

- 1 100kΩ light-dependent resistor (LDR1) **[GL5528]**
- 2 100kΩ 2 10kΩ 2 220Ω 15 68Ω

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shorter as you reduce it and longer as you increase it.

A long press of A will return to the time display, while a long press of B in this mode will switch to adjusting the other end of the range, ie, how much it will dim at low light levels.

Similar to the maximum brightness

setting, you have a fairly wide range of adjustment here, as you may wish the clock to fully dim before it is in total darkness.

Or you may wish for it to never go to minimum brightness, even when the LDR sees no light. The default is somewhere between those two extremes.

While making this adjustment, the LDR is active and the display will dim based on the current ambient light level and the current setting so you can see its effect. To simulate the clock being in darkness, you will need to cover the LDR with something opaque, like a credit card.

Small objects can easily have light leak around the edges, so make sure the object is touching the whole face of the LDR and extends beyond it in all directions if you want to simulate total darkness. Being in a dimly lit room for this adjustment will help.

In this mode, a long press of button A will return to time display mode, while a long press of button B will go back to adjusting the maximum brightness, as described above.

Using it

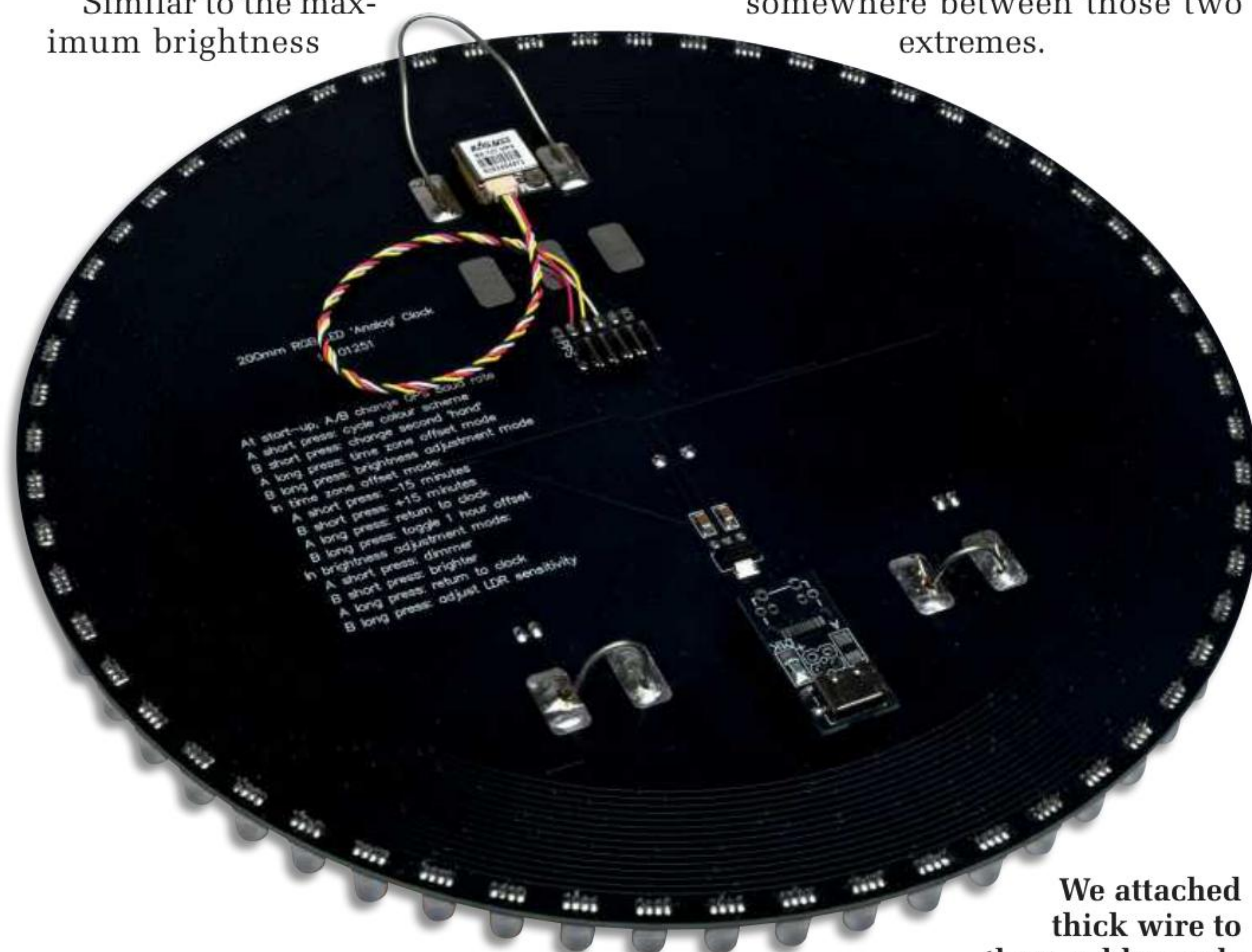
Once you have set your time zone offset, confirmed that your time source is working, adjusted the brightness levels and chosen your preferred colour scheme and mode, the clock is set up and ready for use. If it loses power, when it regains it later, it will go back into exactly the same mode. It just might take a while for your time source to resume, especially if it's a GNSS/GPS module.

All that remains is to hang it on the wall and connect the permanent power supply arrangement. We recommend soldering an inverted-U-shaped piece of tinned copper wire between one pair of pads on either side at the top of the clock. Bend it so that it will comfortably hook around the head of a screw in your wall, or a wall hook.

A couple of smaller loops soldered across the two pairs of pads near the bottom of the clock will stop the bottom of the clock from touching the wall. However, you may wish to have it angled down, as it could make it easier to read. So you could omit those loops, or make them stick out less than the upper one.

If there is no power source under where you want to hang the clock, you could run a thin figure-8 cable from CON1 up behind the clock, then along the wall and then down to a more convenient location. The wire will be less visible if it's the same colour as the wall (you could paint it).

While few homes have picture rails anymore, if you're lucky enough to have them, they are an excellent way to hide such wire runs! **PE**



We attached thick wire to these solder pads on the rear of the Clock, so that it can be hung on a wall.



The clock face is designed so it doesn't take attention away from the LEDs.

USB Power Adaptor

Project by Nicholas Vinen

This simple and cheap PCB provides an easy way to add a USB socket to a 5V DC powered device. It accepts a Type-C or mini/micro Type-B socket and provides optional reverse power flow/reverse polarity protection and LED power indication.

I wanted to add a USB socket to my RGB LED Clock (the article just before this one) for power, but I couldn't fit one near the PCB edge as it is totally occupied with LEDs. Horizontal USB connectors won't work if placed in the middle of the board. Tim Blythman suggested I mount the socket on a small, separate PCB and suggested that it could have other uses. Hence this project.

To make the board as flexible as possible, I have placed four different USB socket footprints on this tiny $12.7 \times 26.5\text{mm}$ board. You can attach the very common SMD mini Type-B socket, one of two readily available micro Type-B sockets or one of two fairly standard Type-C sockets. Use whichever suits your needs.

The resulting 5V DC is available on a set of four pads in the middle of the PCB. These allow a two-way pin header to be fitted in eight orientation and polarity combinations.

This is especially handy if you're mounting this PCB to another one via a pin header, as you can choose which direction the USB socket will face (north, south, east or west) regardless of the header polarity.

You don't have to attach this board to another one; you can solder a pair of wires, or a header and use jumper wires. In fact, it'd be a convenient way to feed 5V to a breadboard from a computer or USB charger.

If using a micro Type-B socket, you can choose one with or without through-hole mounting pins; the PCB will accept either. Having said that, we've specified the type with pins in the parts list as it is easier to mount and more secure once soldered.

Similarly, for USB Type-C, you can use a six-pin power-only socket or a 12-pin power-plus-data socket (the data pins are not connected).

The circuit is very simple, as shown

in Fig.1. You can use a 1A or 3A schottky diode for D1, or a 0Ω resistor (shown as a dashed link shorting D1 out). USB connectors are polarised, so in theory, you don't need D1 for reverse-polarity protection.

Its main purpose is to prevent power from feeding back into the USB power source if the target board is separately powered. So whether you fit D1 or a 0Ω resistor will depend on whether that is possible in your application.

If you're wiring up the USB cable yourself, or if its forward voltage is irrelevant (eg, the target board immediately reduces it to 3.3V with a low-dropout regulator), you may still want to fit D1 for reverse polarity protection.

You don't need to fit the LED and its series resistor if you don't need a power-on indicator. As for the other two resistors, they are only required if you are fitting a USB Type-C socket, to signal to the power source to supply 5V. For Type-B sockets, you can just leave them off.

Construction

Depending on which socket you are using, follow the relevant overlay diagram: Fig.2(a) for USB-C power only, Fig.2(b) for USB-C power and data, Fig.2(c) for mini-B or Fig.2(d) for micro-B. All other required or optional components are shown fitted. If you don't need LED1, leave it and the resistor on the opposite side of the board off. If you don't need D1, replace it with the 0Ω resistor.

Start by fitting the USB socket. It will make soldering easier if you spread a thin layer of flux paste over all the pads for your particular socket before you place it on the board.

Only the USB-C power-only socket lacks locating posts; the others should snap into place and you can then tack

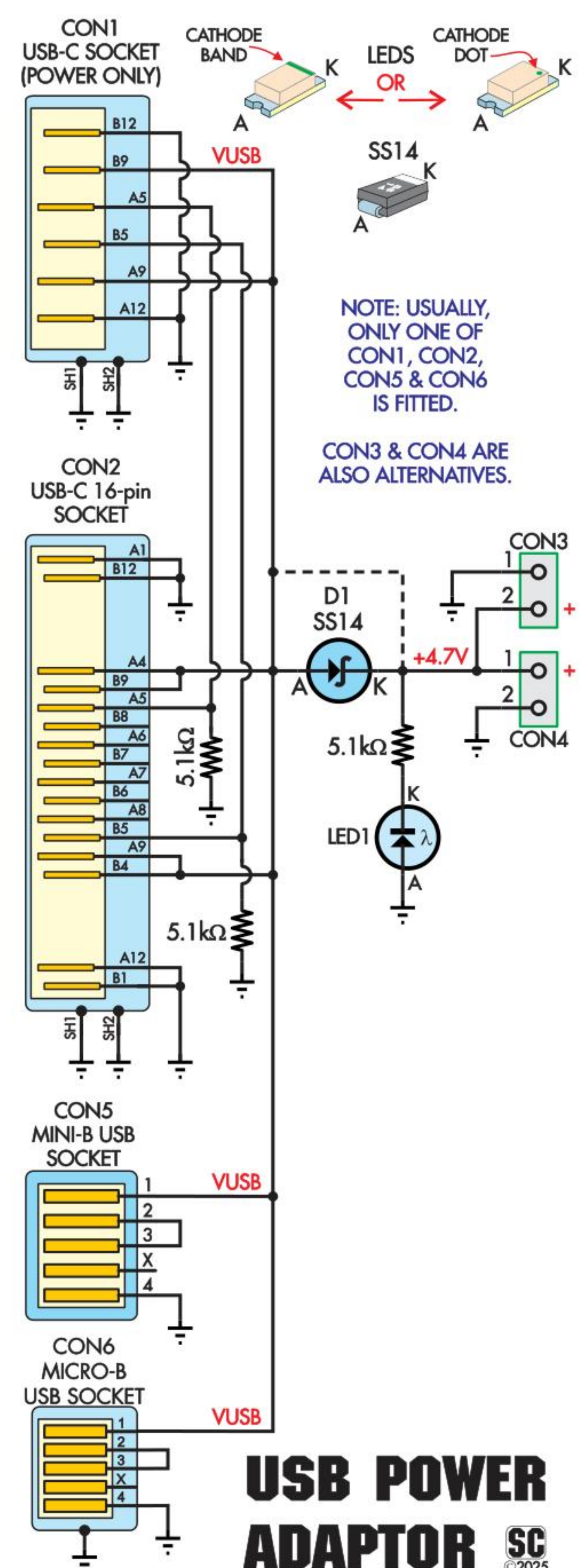


Fig.1: the four possible USB sockets are wired in parallel and it has provision for the two $5.1\text{k}\Omega$ pull-down resistors needed for a Type-C socket to receive 5V. D1 prevents power flowing back into the USB socket, while LED1 provides power-on indication.

Constructional Project



From left-to-right: the Mini USB, USB-C & Micro USB versions.

one pin and check that all the pins are aligned. For the USB-C power-only socket, you'll have to position it by eye initially. Remelt that initial solder joint and nudge it until its six relatively large leads are over their pads.

Add a bit more flux paste over the remaining leads and then solder them. Once those leads have been soldered, you can solder the mechanical mounting pins or tabs. The USB-C power+data socket is designed for 1mm-thick PCBs, which is why we've specified this board that way. Otherwise, its mounting tabs won't go all the way through the board.

You may need to turn your iron up a bit while soldering the mechanical mounting tabs as the USB socket case will draw heat away from them. Turn it back down when you've finished.

Most sockets have pins that are closely spaced, so it's likely you will have some bridges between them now. If you do, add some more flux paste and then press solder-wicking braid down on them with the tip of your soldering iron. Wait for the solder to flow, then slide the braid away from the pins. It should remove the excess solder and leave behind nice-looking joints.

Now is a good time to clean off any flux residue, either with a specialised flux remover, isopropyl alcohol or methylated spirits. Then inspect the board under magnification and good light. Verify that all the USB socket solder joints are good. If not, add some more flux paste and rework them, either by adding more solder or removing excess solder with the wicking braid.

If you fitted either of the USB-C sockets, you'll now need to install the two side-by-side 5.1k Ω resistors. Without them, you may not get power.

Now move on to diode D1. If fit-

ting it, make sure it's orientated as shown. Otherwise, replace it with the 0 Ω resistor, so there is a path for current to flow from the USB socket to CON3/CON4.

If you want the power indicator LED, solder it now. It is also polarised. The best way to do this is to use a DMM set on diode test mode to probe the ends of the LED until it lights up. The black probe will be touching the cathode when it does, so that is the side you solder to the pad marked K on the PCB. If fitting

this LED, don't forget its series resistor; otherwise, it can be left off.

That just leaves pin header CON3/CON4. There are two + symbols shown in two corners; the other corner pads are ground (ie, negative). There are four possible positions that you can solder a two-pin header here, on either side of the board.

Whichever one you choose, one of its pins will go to a pad marked with a + symbol. So choose the location that gives your required polarity (if it matters).

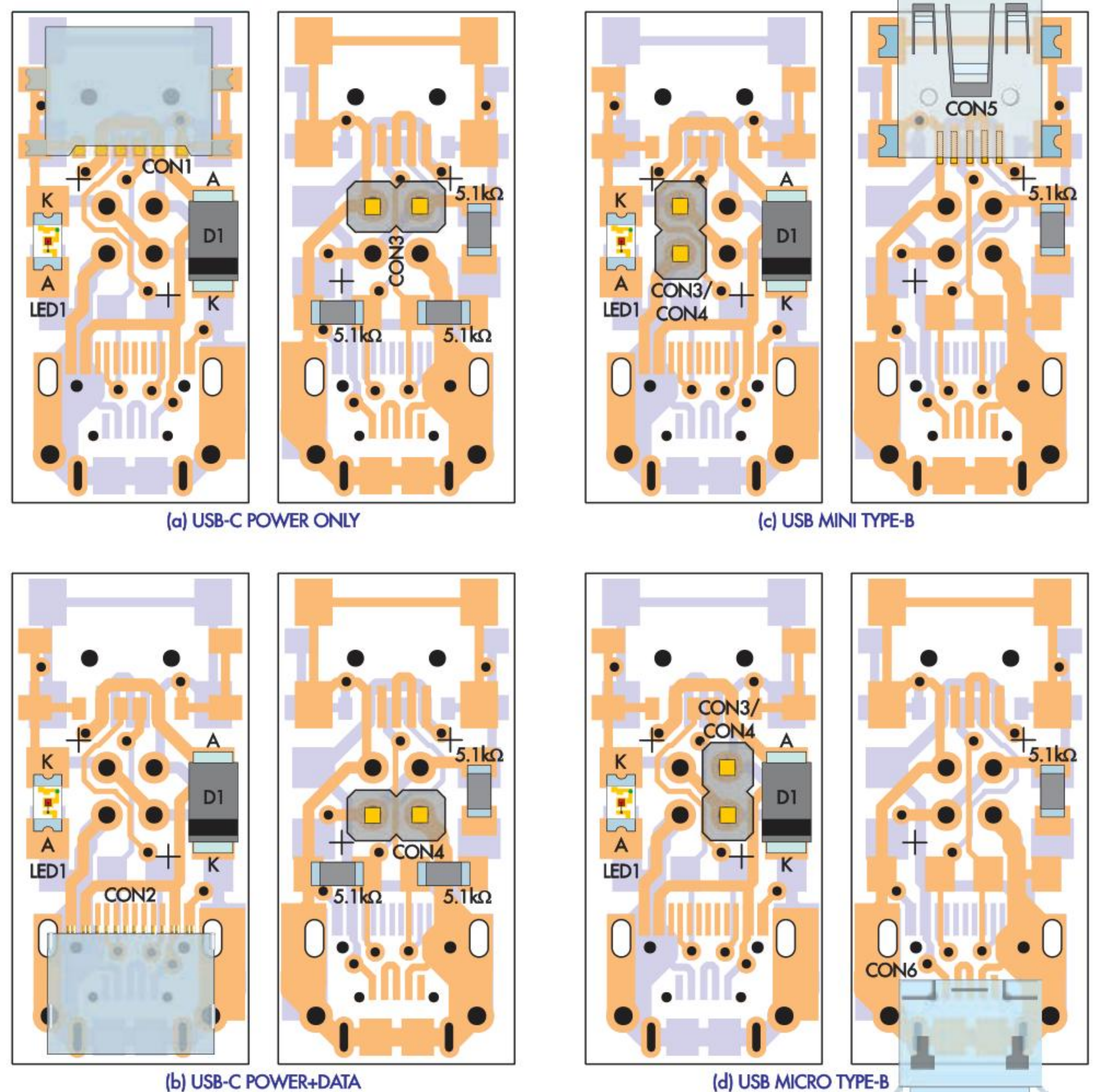


Fig.2: follow the appropriate overlay diagram for the socket you are using. All show D1 and LED1 fitted but you can replace the former with a 0 Ω resistor or wire link, or omit the latter, if you want. The presence and location of CON3/CON4 will also depend on your requirements.

USB Power Adaptor

Parts List – USB Power Adaptor

1 1mm-thick black double-sided PCB coded 18101251, 12.7 × 26.5mm
2 5.1kΩ M3216/1206 SMD resistors (only required for Type-C USB sockets)
1 0Ω M3216/1206 SMD resistor/link

Pick one of these sockets:

1 SMD Type-C USB power-only socket (CON1) [GCT USB4135 or equivalent]
1 SMD Type-C USB 2.0 socket (CON2) [GCT USB4105 or equivalent]
1 SMD mini Type-B socket (CON5) [Molex 0675031020 or equivalent]
1 SMD micro Type-B socket (CON6) [GCT USB3080-30-01-A or equivalent]

Optional parts

1 2-pin header (CON3/CON4)
1 SS14 (1A), SS34 (3A) or equivalent schottky diode, DO-214AC (D1)
1 M3216/1206/SMA SMD LED plus 5.1kΩ M3216/1206 SMD resistor (LED1)

The power + data version of the USB-C Power Adaptor. We didn't have a 0Ω resistor on hand so used a piece of wire instead.

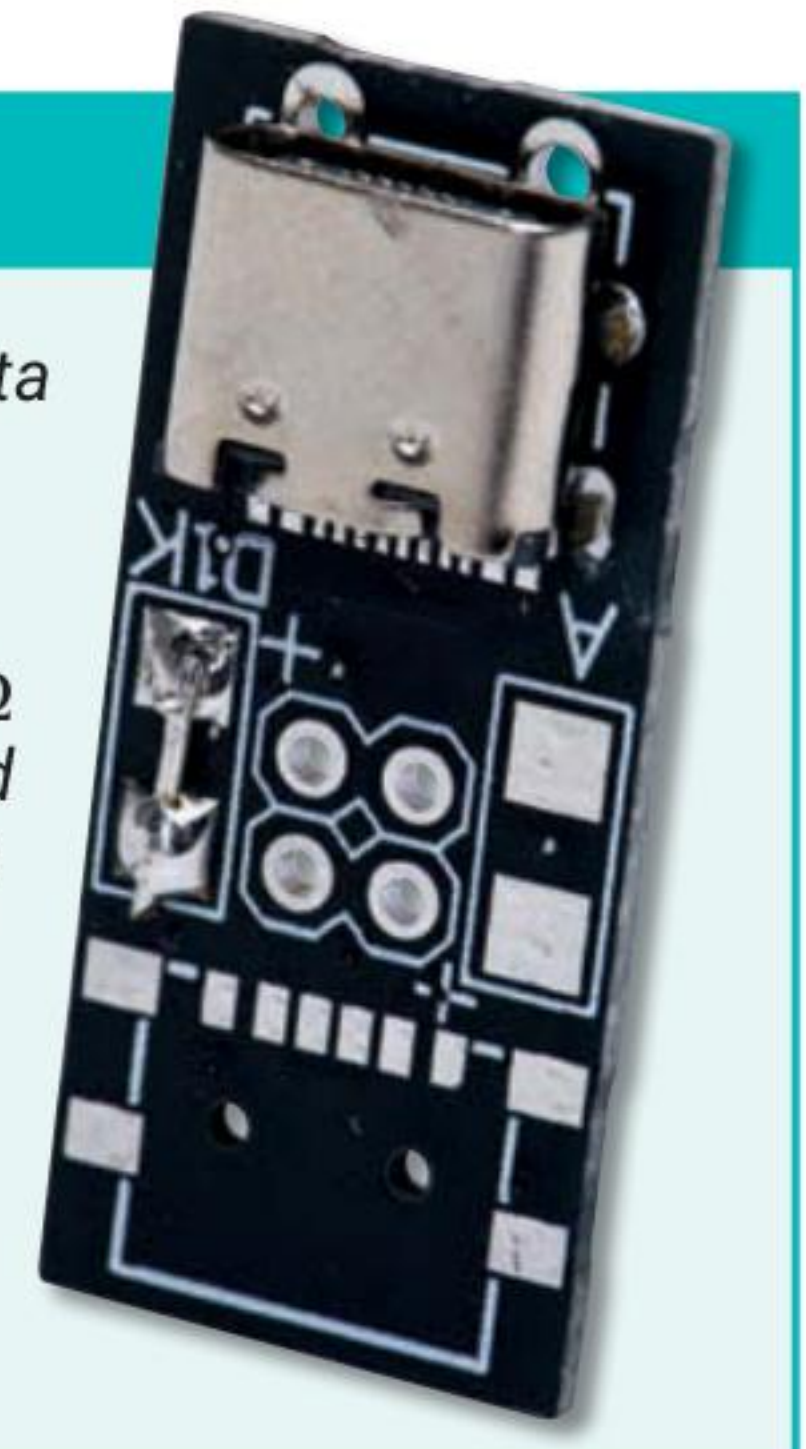


Fig.2 shows four of the possible locations for that header. Alternatively, simply solder two wires to these pads, one to a pad marked + and the other to an unmarked (ground) pad.

The accompanying photo shows the USB Power Adaptor fitted with a Type-C socket mounted on our RGB LED Analog Clock (presented earlier in this issue) using CON3. That position was chosen as it matched the polarity of the power header on the Clock PCB.

Testing

Plug your assembled board into a USB power supply and use a DVM to check the output at CON3/CON4. If you connect the red probe to a + pad and the black probe to one of the other two, you should get a reading of about +5V, or +4.7V if you fitted D1

rather than a 0Ω resistor or wire link.

If you don't get a voltage reading, check that your supply is on and that diode D1 is orientated correctly or linked out.

If you have fitted LED1, it should light up. If you can measure voltage but it isn't on, it may be backwards or have a bad solder joint. Also check the series resistor's solder joints.

All that's left is to wire this up or solder it to your target board, apply power and check that it works. Make sure you get the output connection polarity right!

Note that the final version of the PCB fixes a couple of minor problems with the prototype ones shown in the photos. It's a little bit shorter so the plugs can reach the sockets more easily, the USB-C 2.0 connector has four solder pads to secure the shell

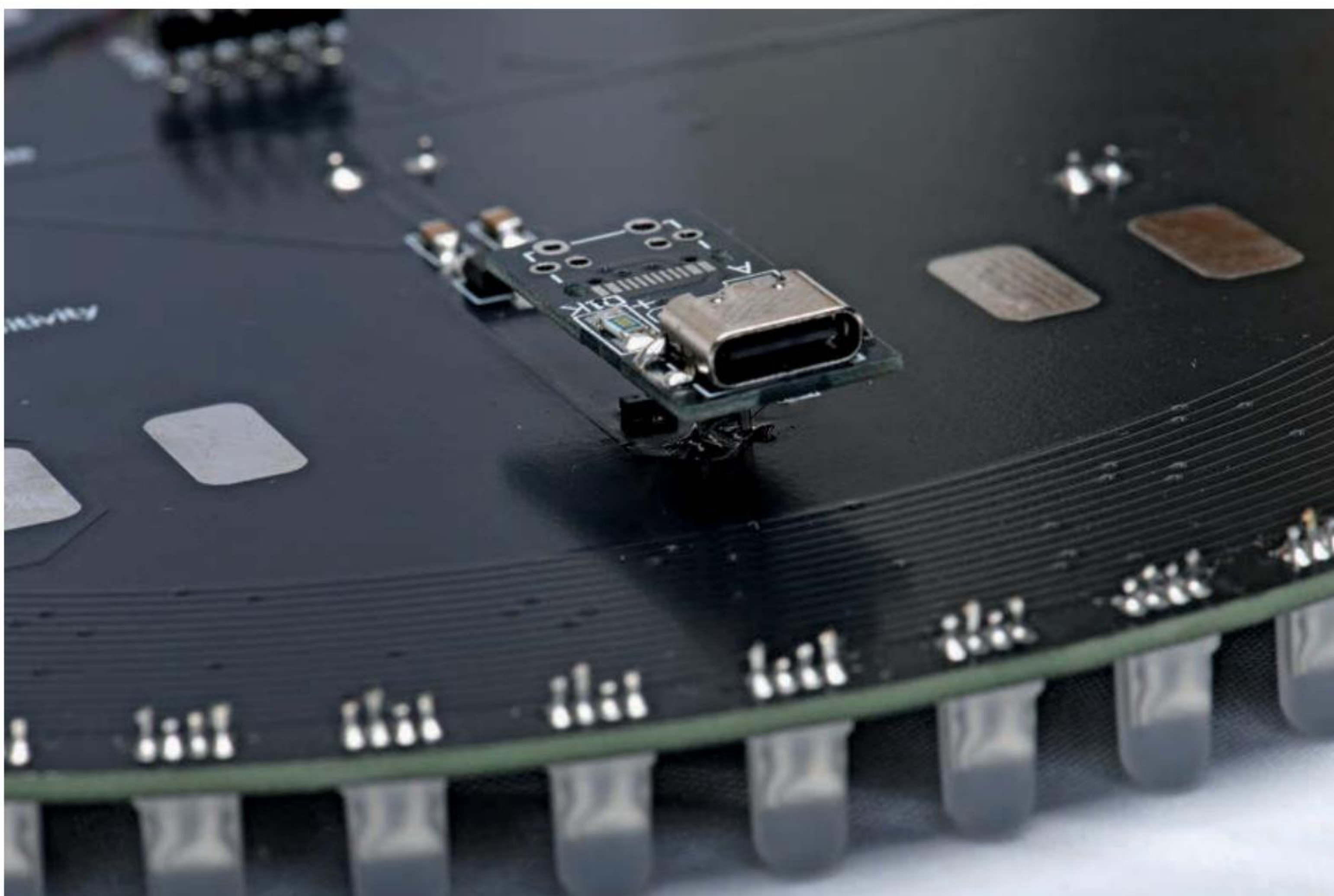
rather than two, and the micro-B footprint was improved to make it easier to solder and more secure.

Using it

Besides the RGB LED Analog Clock, some of our recent projects that this board could potentially be used with include:

- Coin Cell Emulator (November 2024)
- TQFP Programming Adaptors (October 2024)
- Dual-Channel Breadboard PSU (December 2023)

In some cases, the connection would be made via the ICSP (in-circuit serial programming) header, which has VDD and GND pins next to each other, suitable for connection to CON3/CON4 on this board. **PE**



One of the USB-C versions of the Power Adaptor attached to our new RGB LED Clock.

USB Power Adaptor Kit (SC7433, ~£5 + P&P) from SILICON CHIP (sent from Australia)

Includes everything in the parts list; choice of one USB socket:

1. USB-C power only
2. USB-C power+data
3. mini Type-B
4. micro Type-B.

<https://siliconchip.au/Shop/20/7433>

Circuit Surgery

Regular clinic by Ian Bell

Measuring circuit frequency response using a PC sound card, part 9: differential signals and the THAT1646 IC

Over the past few months, we have been looking at circuit frequency responses, including implementing a measurement system using the ‘sound card’ (audio interface) of a PC or laptop together with a free app called REW (Room EQ Wizard).

This was motivated by the need to measure the response of a practical implementation of an example digital filter (from our series on DSP) without using advanced lab test equipment.

In the November and December 2025 issues, we described parts of the signal conditioning circuitry that might be needed to interface the circuit under test to the sound card’s line inputs and outputs (see Fig.1). Since then, we have covered various aspects of circuit theory related to frequency response and simulation techniques in LTspice to help us investigate this.

Parts of the interface circuit may need to drive large capacitive loads, which can result in instability. This topic is closely related to frequency response analysis, while also being applicable well beyond the specifics of the sound card interface. The most recent articles have covered capacitive loading of op amps and techniques to mitigate the instability this can cause.

The output from the sound-card-based measurement circuit to the computer’s audio interface was provided by a THAT1646 balanced audio line driver IC. However, we have not discussed this part of the system in detail until now.

Balanced lines and differential signals use two conductors (in addition to ground). They are important and widely used in audio systems, as well as other electronic systems. We will mainly cover the basics of differential signals this month. As I ran into problems simulating the THAT1646 IC, we

will take a sidestep to look at how to deal with them.

As with our discussion of op amp stability, these topics have wider applicability than just the frequency response measurement system.

The THAT1646 and similar devices are better than typical op amps at handling capacitive loads, such as long audio cables. This was one of the reasons I used it, along with its capability of driving both single-ended and balanced audio inputs.

Desktop PC audio interfaces (sound cards) typically provide standard consumer stereo audio line inputs, which are ideal for use with the measurement system. The system uses two signal paths, one for measurement and one for reference, so two inputs are required (see Fig.2).

The signals from the measurement system are not in stereo, but that does not matter – they simply provide two streams of data that are interpreted as required by the measurement software.

Laptops often only have a 3.5mm TRRS (tip, ring, ring, sleeve) headset connector, which provides stereo outputs and a mono microphone input. With this type of interface, it is not possible to feed in the two analog signals required by the measurement system. Furthermore, the single input is aimed at microphones, rather than audio line levels.

An external audio interface connected by USB is required if the two-channel measurement process is to be used with such computers.

The cheapest USB devices typically often only provide mono microphone inputs, but some provide two line-level inputs and two line-level outputs, similar to a PC. Some of these are aimed at home recording.

Starting at a slightly highest cost, there

are audio interfaces aimed at musicians and other professional audio applications. These often provide balanced (differential signal) inputs and outputs, which are common in professional audio.

I have one of these (not a high-end one, unfortunately!) so another reason I used the THAT1646 was to experiment with using this unit, as well as the line inputs on my PC’s built-in interface.

The THAT1646 line driver

The THAT1646 is an audio differential line driver IC that claims improved performance over pin-compatible devices like the DRV134 and DRV135 from Texas Instruments and the SSM2142 from Analog Devices. This is due to a design approach that the manufacturer has trademarked as “OutSmarts”. It is a feedback-controlled balanced output driver that behaves properly when the load isn’t perfectly balanced, especially during clipping.

Douglas Self described how to do something similar using op amps in his Small Signal Audio Design book (Chapter 15 [Line Outputs], pp386-387, “Quasi-Floating Outputs”).

The THAT1646 IC is produced by THAT Corporation, a Massachusetts-headquartered IC manufacturer focusing on professional audio devices such as microphone preamplifiers, voltage-controlled amplifiers, line drivers, receivers and digital signal processors (DSPs).

The THAT1646 is available in DIP-8, SO-8 and wide SO-16 packages. Fig.3 shows the internal circuit structure and

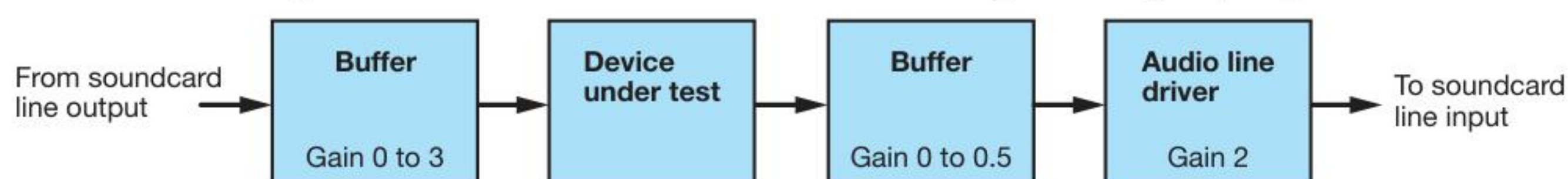


Fig.1: signal conditioning circuitry for sound card frequency response measurement.

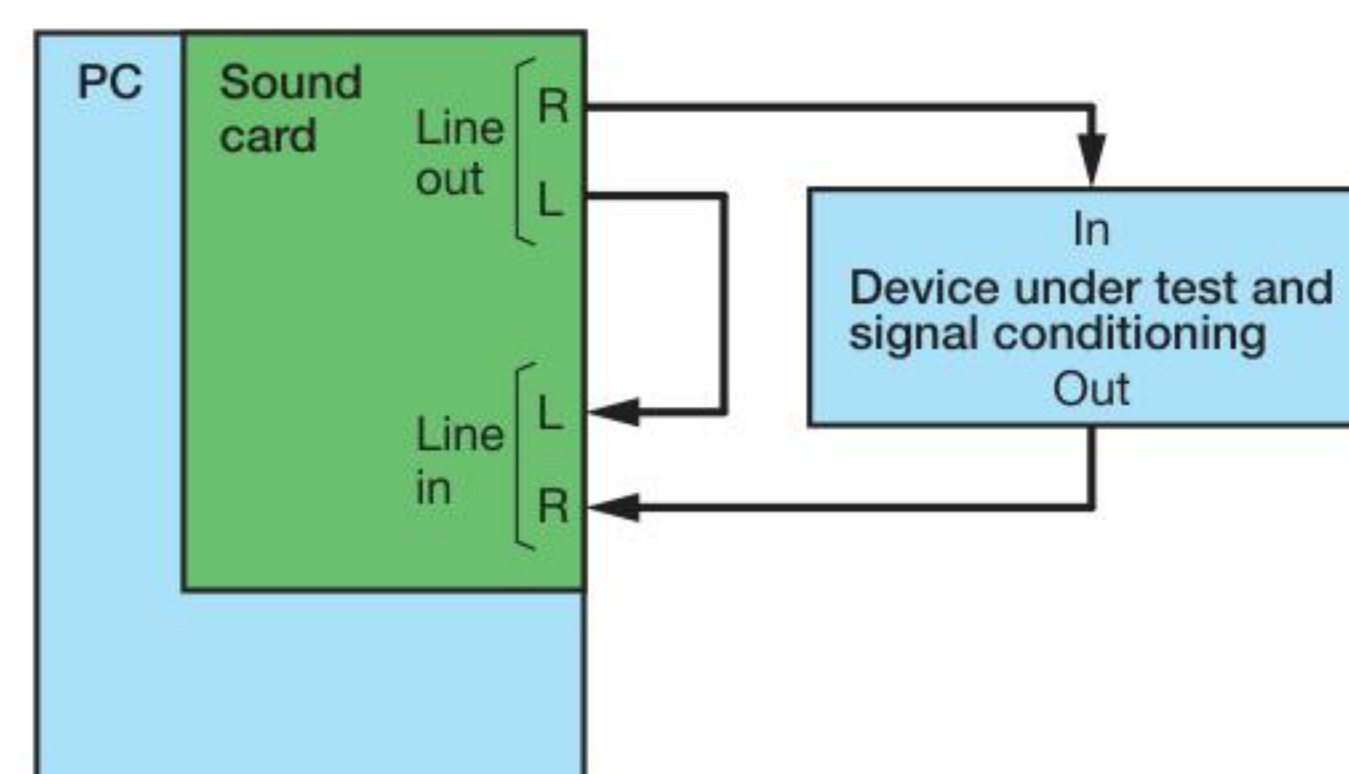


Fig.2: the setup for testing an electronic device using a sound card.

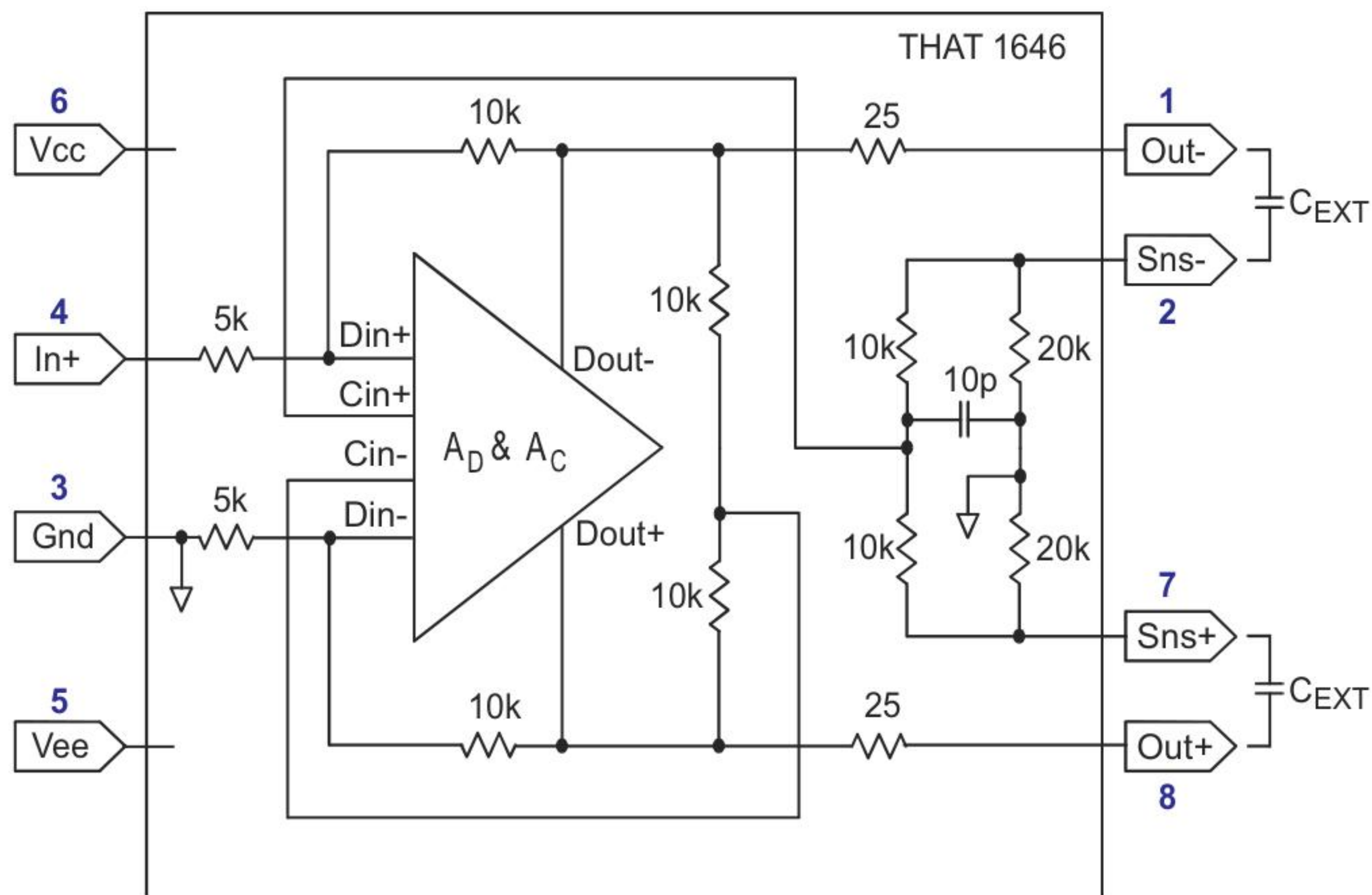


Fig.3: the circuit structure and pinout of the THAT1646 (from the device data sheet).

pinout for the DIP-8 and SO-8 versions. The core of the circuit is a fully differential amplifier. The resistors and capacitors form the feedback network, which is significantly more complex than a basic op amp amplifier.

The op amps we have investigated in detail recently have differential inputs and a single output. They amplify the voltage difference between their inverting and non-inverting inputs and provide the amplified signal on their single output.

The amplifier in the THAT1646 has a differential input like an op amp (Din+ and Din- in Fig.3) but also has a differential output – the output signal is between the two outputs (Dout+ and Dout-) rather than from a single ground-referenced output.

A close look at Fig.3 will reveal that, although the internal circuitry uses a fully differential amplifier, and the output is differential, one of its 5kΩ input resistors is internally grounded, so the input to the THAT1646 is single-ended. A similar device, the THAT1606, has both differential inputs and outputs.

The 5kΩ resistors at the inputs (along with the 10kΩ feedback resistors from Dout to Din) play a similar role to the gain-setting resistors in an inverting op amp amplifier.

Whenever we have a differential signal between two points in a circuit, we can also have a common-mode component, which is the same at both points relative to ground (the average of the two signals). In a differential amplifier, the DC bias that keeps the transistors at their correct operating point is typically common-mode and needs to be stabilised using common-mode feedback.

The amplifier in the THAT1646 has separate inputs for common-mode feedback (Cin- and Cin+).

Differential & balanced signals

Before discussing the device in more detail, we will look at the basics of differential signals and balanced connections. These terms can cause confusion and are sometimes incorrectly used interchangeably, particularly in audio; however, the terms do not have the same meaning.

Both differential signals and balanced connections are used to improve signal integrity. Common uses include professional analog audio and low-voltage differential signalling (LVDS) in high-speed digital systems.

Fig.4 shows scenarios where two systems are connected together with a driver sending a signal to a receiver. This is a generic situation; the signals could be analog or digital and carry any kind of information. The upper diagram shows a single-ended connection, which is inherently unbalanced. The lower diagram could represent a balanced and/or differential connection.

The single-ended connection is the most basic direct electrical output-to-input connection between components or systems. There is single signal conductor that carries a signal voltage (v_s) referenced to ground.

There also needs to be a good electrical connection between the driver and receiver grounds, as both ends must use the same reference for v_s to be interpreted equally at both ends of the connection.

Any current flow from driver to receiver must return along the ground connection. The ground connection could be various things, such as the ground plane on the PCB, a ground track in an IC, or a conductor in a cable, including the shield of a shielded cable.

Both differential signals and balanced connections use two wires other than ground, so we can observe the voltage on each wire individually with respect

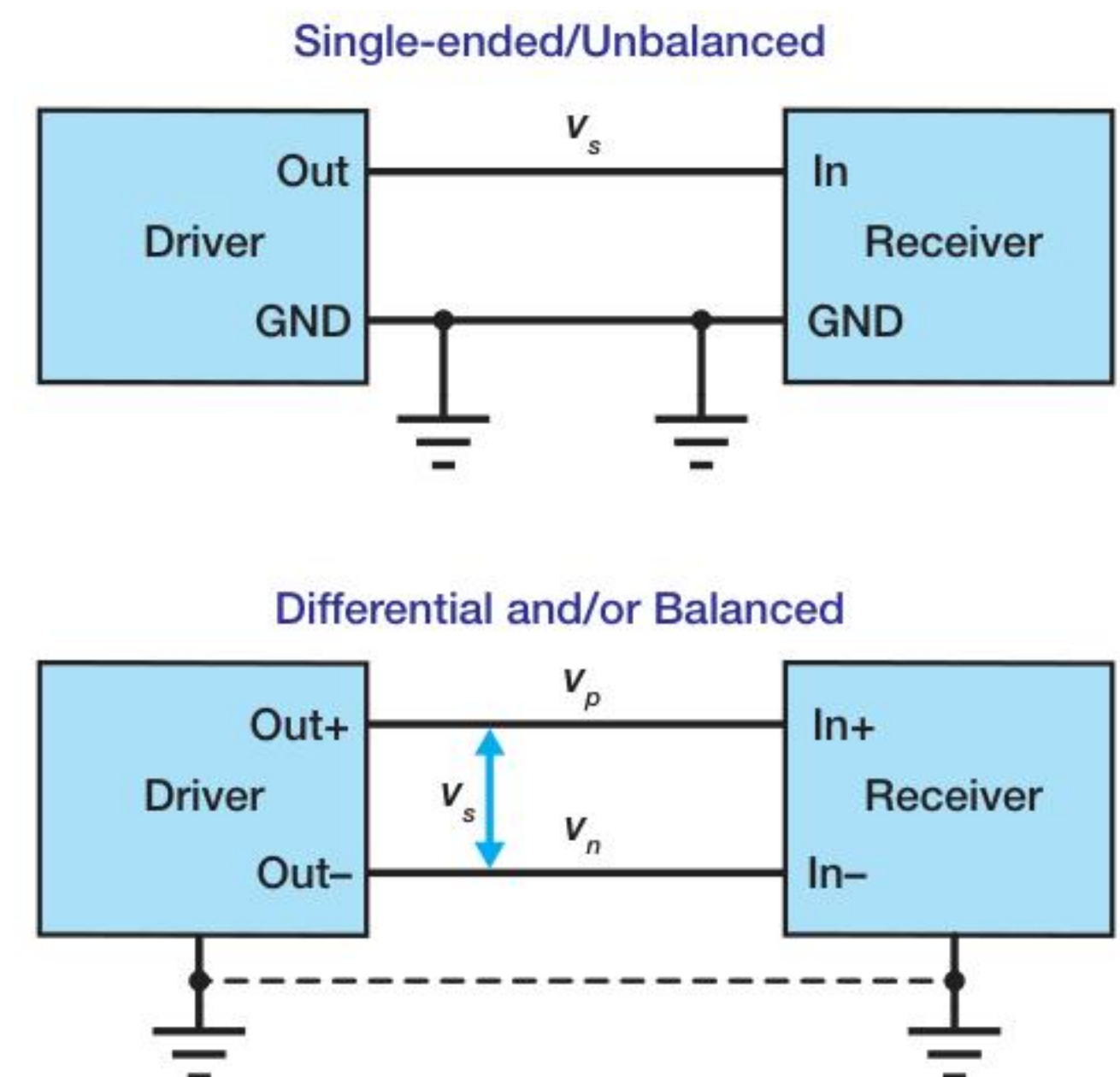


Fig.4: connection types and signals.

to ground. The actual signal is equal to the difference in voltage between the two wires. If the two voltages on the two wires are v_p and v_n the signal is $v_s = (v_p - v_n)$.

In Fig.4, we labelled the differential signals or balanced connections as positive (P) and negative (N). In audio systems, these are often referred to as “hot” and “cold”.

The signal is differential if it is symmetrical, that is, equal and opposite voltages on the two connections (so $v_p = -v_n$). There are advantages to this, but it is not essential, and one of the individual signals may be zero.

The connection is balanced if the impedances are the same for both conductors. This is sometimes just stated as the impedance to ground, but all impedances (including the driver's source impedance) and the impedance of the conductors themselves must be equal.

The latter is important for longer connections where the total resistance or capacitance (for example) of an audio cable may be significant. In some situations, like very high-speed digital connections on a PCB, the characteristic impedance of the connection as a transmission line must be considered. This is not relevant to audio in normal domestic, studio, and live scenarios.

You can have a differential signal on an unbalanced two-conductor connection. Similarly, signals on balanced connections do not have to be differential. The receiver for a differential signal or balanced connection must have a differential input. The greatest advantage in terms of signal quality will usually be obtained by using differential signals on a balanced connection.

Having the signal voltage between two wires means that it is self-referencing – a ground connection between driver and receiver is not required to obtain an accurate voltage transfer. Also, current from one connection can return via the other.

Transformers have long been used to provide balanced, electrically isolated

audio inputs and outputs (see Fig.5). However, modern driver and receiver ICs can provide high-quality balanced interconnections that are smaller and cheaper, usually with better signal quality.

This is due to the complex frequency-dependent behaviour of transformers, which may influence characteristics such as distortion and transient response. An advantage of transformers is that they provide full electrical isolation, which is important in some situations.

If the signals are not galvanically isolated (no direct conduction path, eg, via transformers or opto-isolators), a ground connection is required between driver and receiver (shown dashed in Fig.4). This is to prevent possible ground voltage differences between the systems, taking circuits out of their operating range or even causing damage.

Grounding issues between connected equipment can also lead to safety concerns if systems are faulty or safe wiring practices have not been followed.

Floating outputs

Another concept relevant to balanced connections is that of a floating output. Often this means a transformer output that provides galvanic isolation, but another important aspect of floating outputs concerns what happens if you short one output to ground.

Although such shorts will not happen in circuits like LVDS connections in digital systems (unless there is a fault), it is not uncommon for balanced audio outputs to be connected to single-ended inputs, resulting in one output being grounded. Not all balanced/differential outputs will tolerate this, particularly if they are not ‘floating’.

If we have a 1V peak-to-peak differential floating output from a transformer and we ground one of the transformer secondary terminals, we will still get a 1V signal. Not all differential-output amplifiers will behave this way, and assuming they can cope with the output short circuit, may be reduced to delivering 0.5V peak-to-peak if one output is shorted to ground.

The THAT1646 does not do this; it will maintain the 1V output if one output is shorted.

Furthermore, some differential-output amplifiers will experience excessive currents if driven into clipping with one output shorted. Again, the THAT1646 does not suffer from this problem, and so is marketed as better mimicking a true floating balanced output than some other line drivers.

LTspice sources

A simple way to create a differential

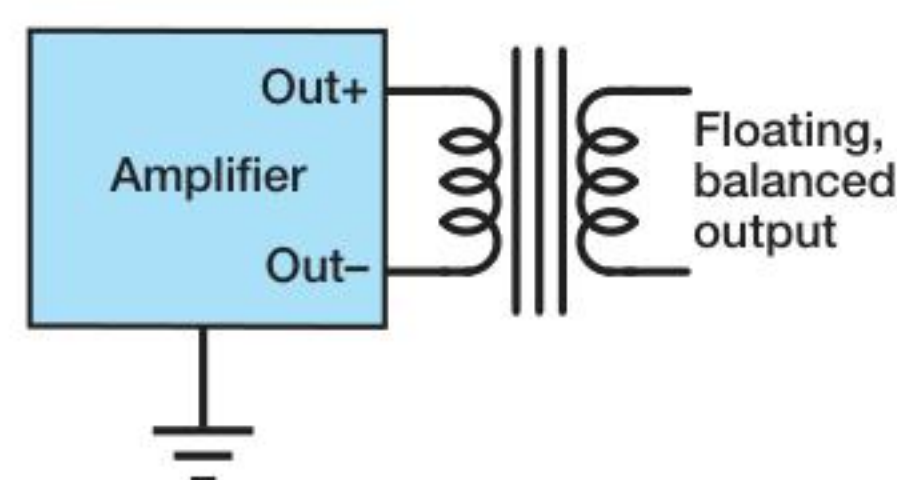


Fig.5: a transformer-coupled output.

signal source in LTspice is to use a floating voltage source, as shown in Fig.6. The “Differential-input circuit” is just a graphic element on the schematic; there is no ground connection for source V1, so the P and N signals are floating.

Unfortunately, due to the way the maths works in a SPICE simulation, the circuit in Fig.6 cannot be simulated. If you try, you get an error message:

This circuit does not have a conduction path to ground! Please flag a node as ground.

The reason for this is that the voltage on nets P and N with respect to ground is not defined; only the voltage between P and N is set by V1. Therefore, LTspice cannot solve an equation to find $V(p)$ and $V(n)$, and as it cannot do this, it cannot run the simulation.

This circuit in Fig.6 is abstract, as it just contains the ideal source; if we included the components for the differential-input circuit, we might create a conducting path to ground and the simulation would run. However, this is not guaranteed, as some common input circuits, such as a transformer primary or the LED in an opto-isolator would not do this.

A possible solution is to add two very high-value resistors from the source output to ground, as shown in Fig.7.

This allows us to simulate the source

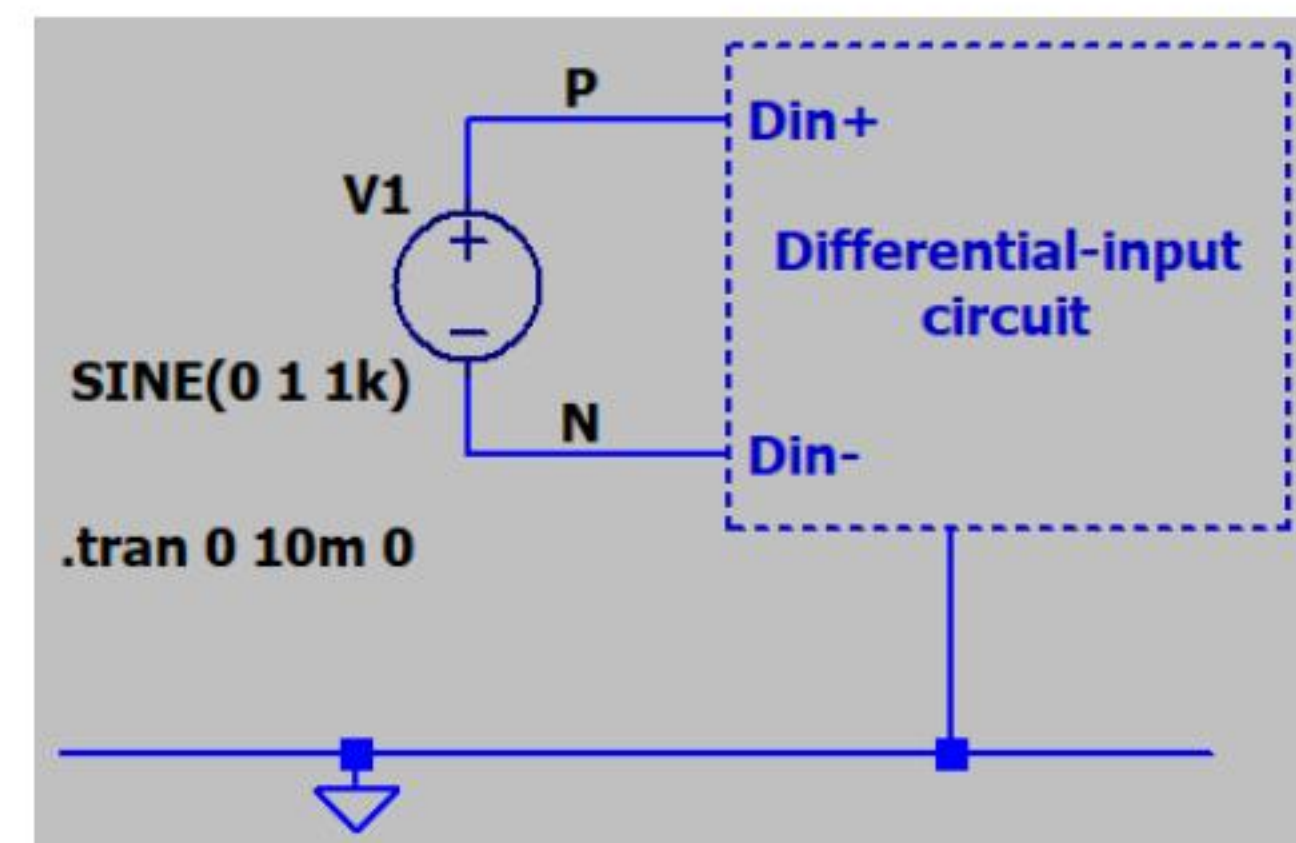


Fig.6: differential signal from a floating source in LTspice.

on its own, giving the results shown in Fig.8. The V1 source produces a differential signal that is a 1kHz sinewave with a peak voltage of 1V (2V peak-to-peak, which is shown in the upper plot). To display the differential signal, we plot $v_P - v_N$ using the LTspice formula $V(p) - V(n)$.

Formula-based waveforms can be displayed via the “Add traces” item of the waveform window right-click menu. You can also display the differential signal via the schematic by first right-clicking on the N net and selecting “Mark Reference”. This will place a black probe symbol next to the net, as shown in Fig.7.

Subsequent left-clicks on other nets will display the voltage on that net relative to the marked reference net. Clicking on the P net will display the differential signal, labelled as $V(P,N)$ (voltage on P relative to N). You can also enter this format manually via

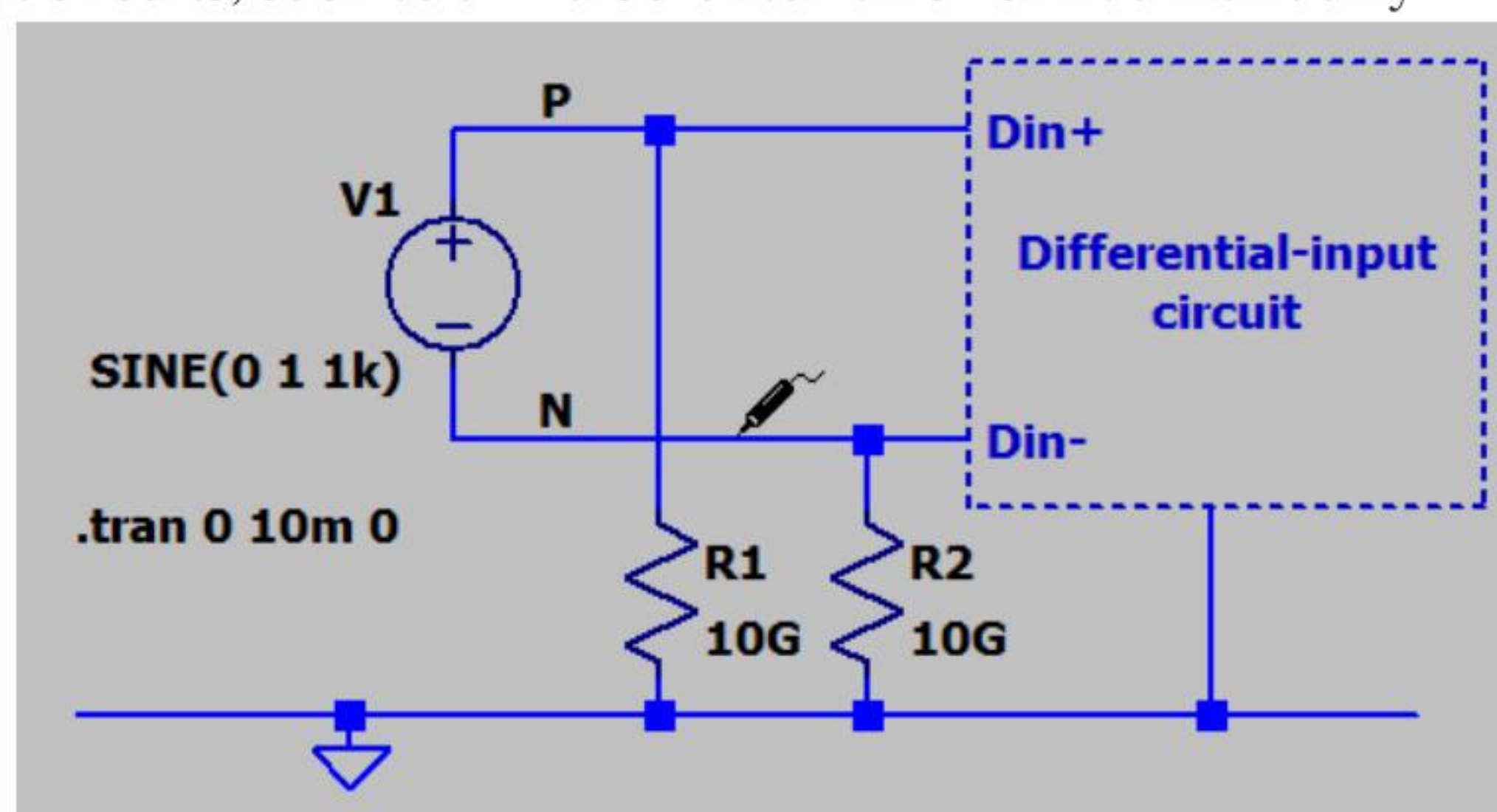


Fig.7: added resistors provide a path to ground.

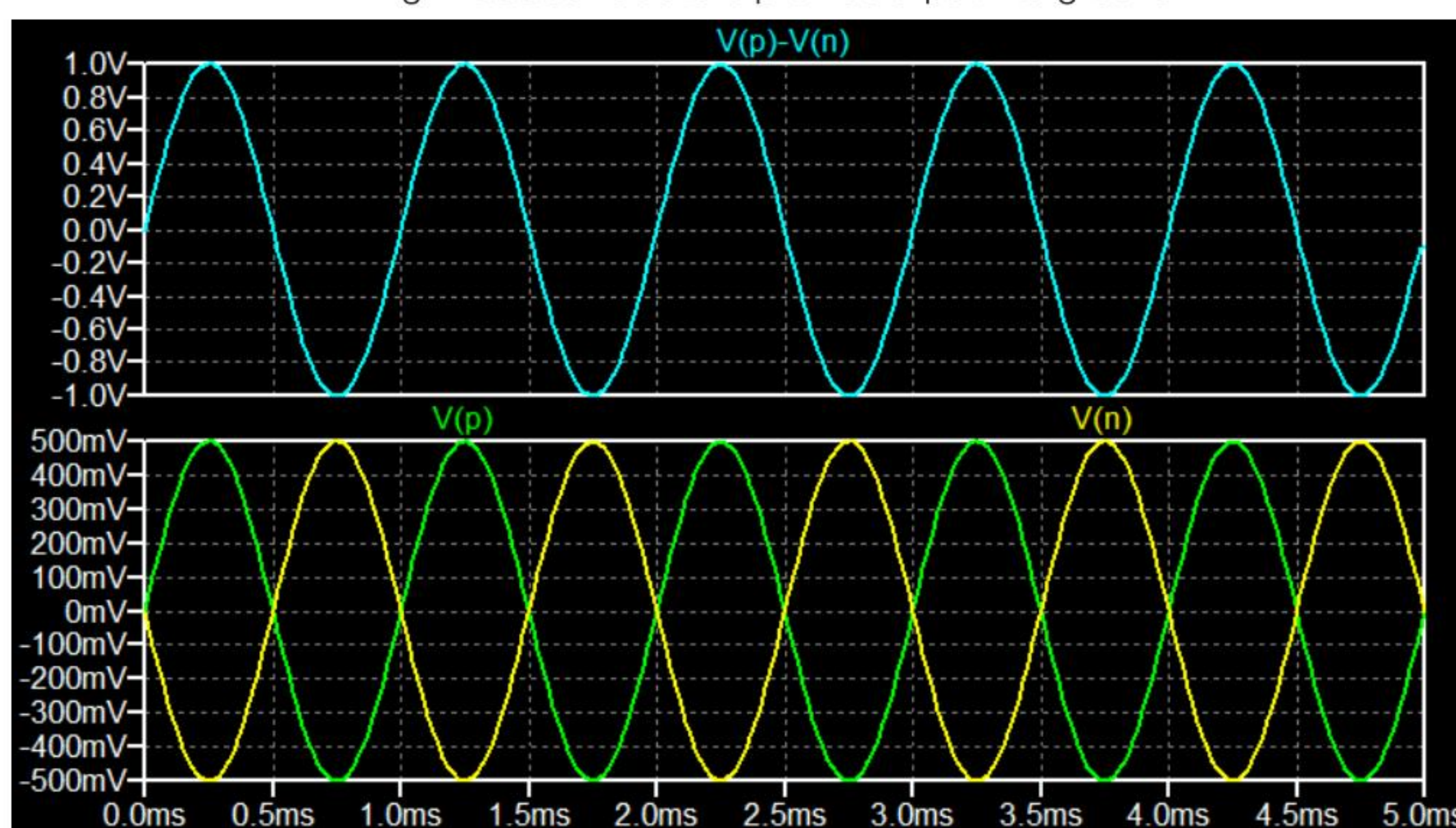


Fig.8: simulation results for the circuit in Fig.7.

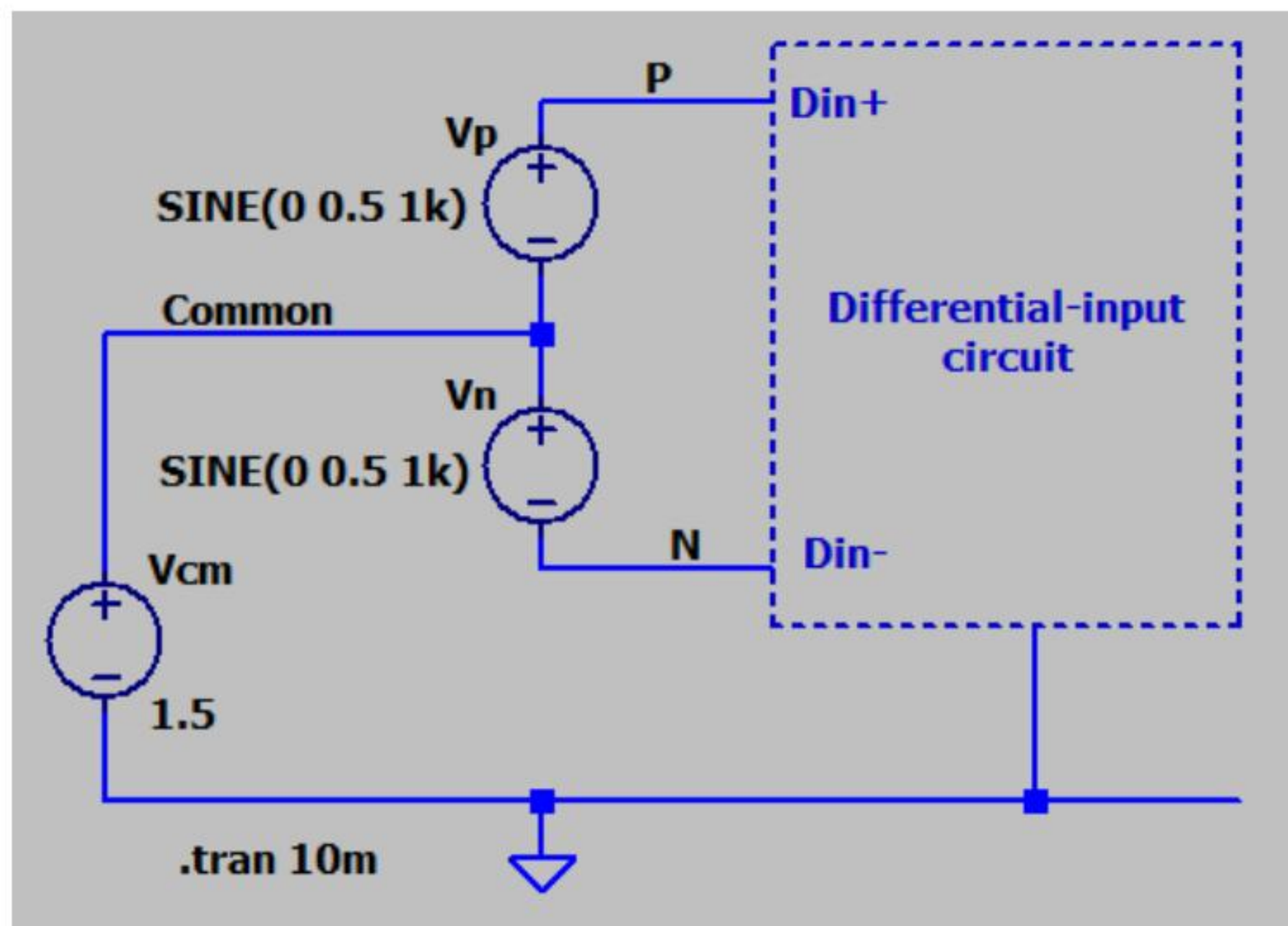


Fig.9: creating a differential + common mode signal in LTspice.

“Add traces” rather than the subtraction formula.

Hitting the escape key will remove the reference probe – make sure you do this if you need to display a normal ground-referenced signal after using the reference probe.

The two individual voltages (v_p and v_n) on the two nets that carry the signal are shown on the lower plot – these are equal and opposite and have peak voltages of 500mV. Because they are opposite, the peak difference between them is 1V ($2 \times 500\text{mV}$), which is the amplitude of the differential signal.

Common-mode components

As mentioned above, differential signals also have a common-mode component. The common-mode is not part of the wanted signal – it cancels out when taking the difference, but it may be important for circuit operation.

For example, standard op amps may fail to operate correctly if presented with too high a common-mode input, even if the differential input signal is within the normal operating range. This

is because the transistors in the circuit can be taken out of their correct bias conditions.

DC common-mode voltages, where present in differential circuits, often relate to bias conditions, so they have to be correct for proper operation. However, in differential interconnections, a common-mode component is often due to non-ideal characteristics or external interference.

The differential signal in the circuit in Fig.7 has a common-mode component of zero – we see that the two individual waves in Fig.8 are centred on 0V. Using a single source like this does not allow us to create a common voltage other than zero. We could add a DC offset to the sinewave from **V1**, but this would be a differential DC voltage, not a common-mode one.

The circuit in Fig.9 has a differential voltage between nets **P** and **N** equal to the sum of the two sources, **Vp** and **Vn**. Each source must produce the same waveform at half the amplitude of the required differential signal. In this case, the two sources have an amplitude of 0.5V, so they produce a 1V, 1kHz sinewave

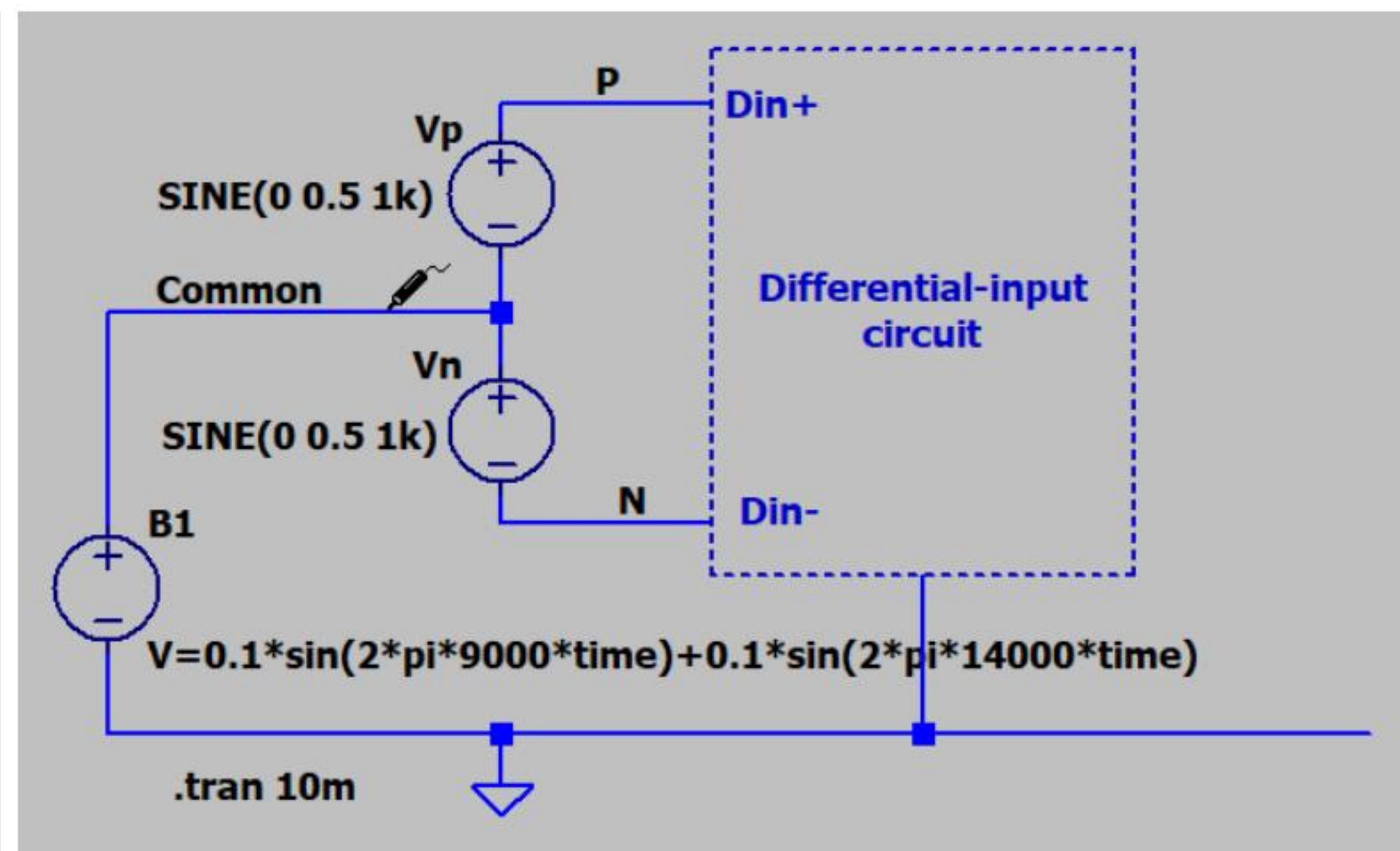


Fig.11: the circuit in Fig.9 with and AC common mode signals.

differential signal, which is the same as that from the **V1** source in Fig.7.

The **Vcm** source produces the common-mode voltage, which is set to 1.5V DC. The simulation results are shown in Fig.10. The upper plot plane shows the differential signal, which is exactly the same as the differential signal in Fig.7.

The lower plot plane shows the individual signals. These have the same amplitude and equal-and-opposite relationship as the corresponding signals in Fig.7, but are offset by the 1.5V DC common-mode voltage. If you take the values of **Vp** and **Vn** at any instant in time and find the average, it will be 1.5V – the common-mode voltage.

We don’t have to use three voltage sources to create a differential signal with a non-zero common-mode component. We can use two sources; one for each of the differential signal nets. These need to be set to $v_{CM} + 0.5v_D$ and $v_{CM} - 0.5v_D$, where v_{CM} is the common-mode signal and v_D is the differential signal.

The advantage of the arrangement in Fig.9 is that it provides a clearer visual indication of the source of the common-mode part, which can be changed by editing just one source.

Common-mode noise

Common-mode voltages do not have to be DC – they can be any signal form or wave shape. Fig.11 shows a version of the circuit in Fig.9 with a behavioural source (**B1**) replacing the DC common-mode voltage. The **B1** source generates a dual sine with frequencies of 9kHz and 14kHz, both with 0.1V amplitudes. The differential signal is the same as in Fig.9 and Fig.7.

The dual sinewave is achieved using the LTspice expression $V=0.1*\sin(2*\pi*9000*time)+0.1*\sin(2*\pi*14000*time)$ for the source. These sinewaves use the standard formula $v(t) = \sin(2\pi ft)$ for a sinewave voltage (v) as a function of time (t) at a frequency of f hertz with no phase shift. The LTspice keyword **time** represents the current simulation time in the expression.

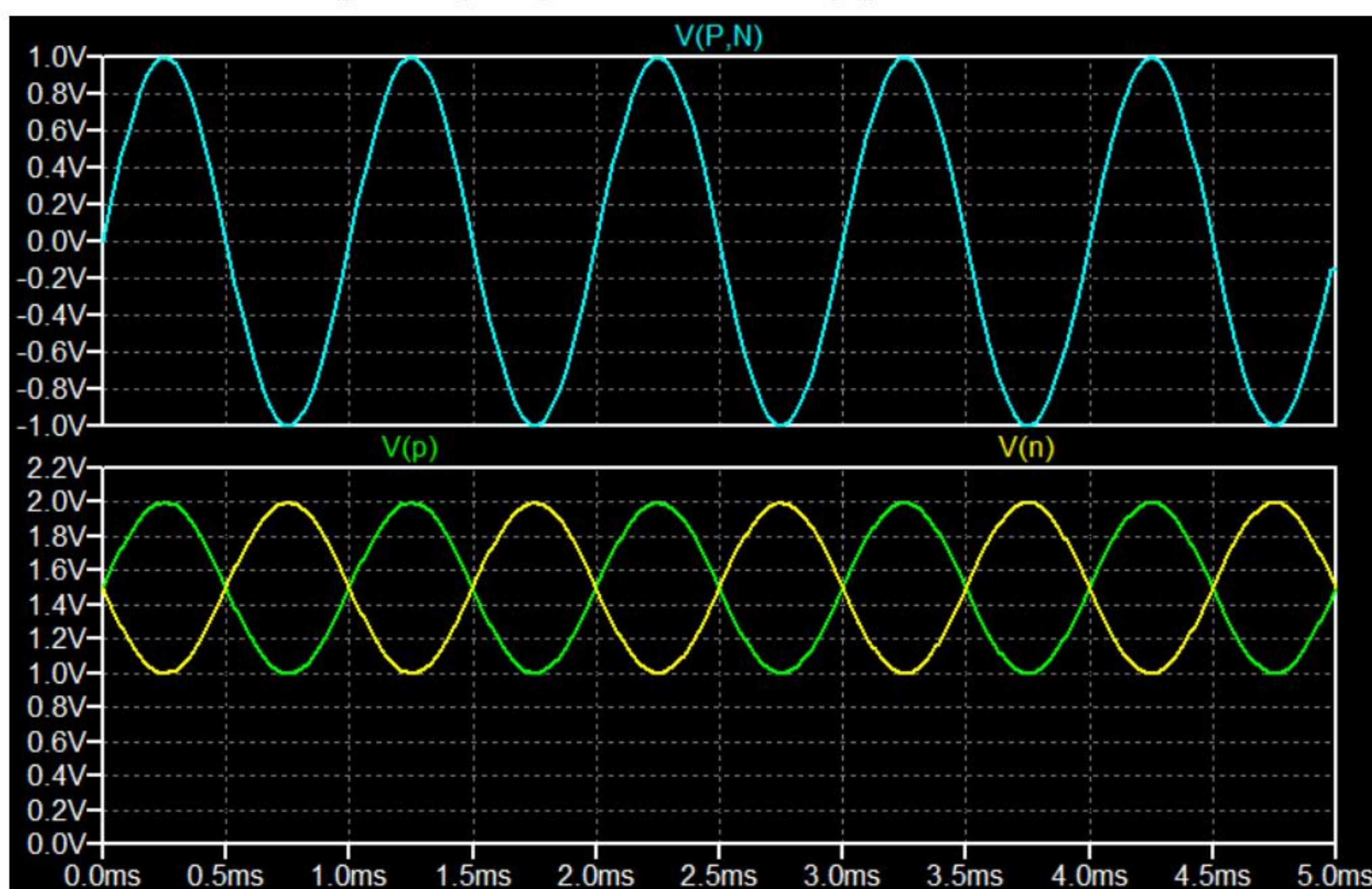


Fig.10: waveforms for the circuit in Fig.9. The differential signal is the same as in Fig.8, but it has different common mode components.

Fig.12 shows the results from Fig.11. The upper plane shows that the differential signal is the same as in the previous examples, although we have zoomed the time axis to facilitate seeing the higher-frequency common-mode signals.

The individual voltages on the two nets carrying the differential signal, measured with respect to ground (**Vp** and **Vn**, green and yellow in the lower pane) look very different from the clean 1kHz sinewave differential signal (cyan, upper pane). The higher frequency signal could be noise induced onto the wires carrying the differential signal.

The lower pane also shows **Vp** and **Vn** measured with respect to the **common** net (grey trace for **Vp** and brown trace for **Vn**). We see that the higher frequency common-mode signal moves in the same direction relative to the two halves of the differential signal and hence is ignored or cancelled out when the difference is taken.

Fig.13 shows a plot of the common-mode signal. In the upper plane (red trace) is obtained by calculating the average of the signal on the two differential signal nets using the formula $(V(p)+v(n))/2$. In the lower plane is plotted directly as the signal on the **common** net (magenta trace).

Noise immunity

The preceding discussion shows that if we have two non-grounded wires carrying a signal between them into a differential amplifier, we have the potential to remove (cancel) any errors or interference that manifests as a common-mode signal.

If this error is δ then the voltage on wire 1 will become $v_1+\delta$ and the voltage on wire 2 will become $v_2+\delta$. The signal is the difference between the two wires, ie, $(v_1+\delta) - (v_2+\delta) = (v_1 - v_2)$, which is the same as the difference signal without the induced noise.

The individual wanted signals do not have to be symmetrical (as in a differential signal) for this work; the key thing is that the error or disturbance is the same on both connections.

Electromagnetic interference is a common problem in electronic systems. An example of where this might occur is with a signal that is carried on long wires (eg, an audio cable) through an electrically noisy environment.

If the interference is in an entirely different frequency range to the wanted signal, it may be straightforward to remove by filtering (assuming you know what filter to apply). However, if the frequencies overlap, as is often the case (eg, 50-60Hz mains and its harmonics are within the audio frequency range), differential processing is very useful.

Electromagnetic interference is effectively a signal source coupled into the circuit or cable capacitively or

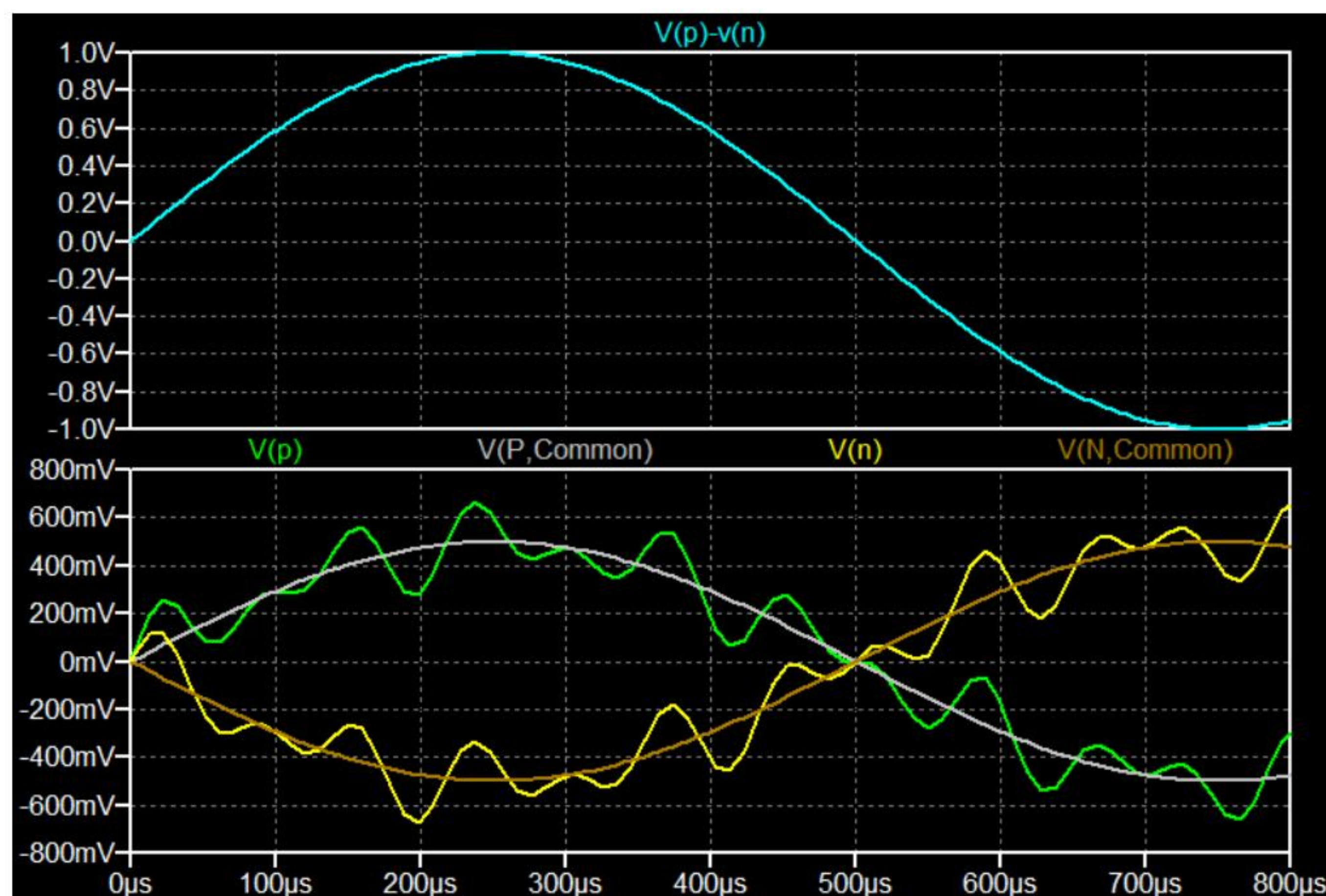


Fig.12: the simulation waveforms for the circuit in Fig.11. This shows the higher-frequency common mode signals.

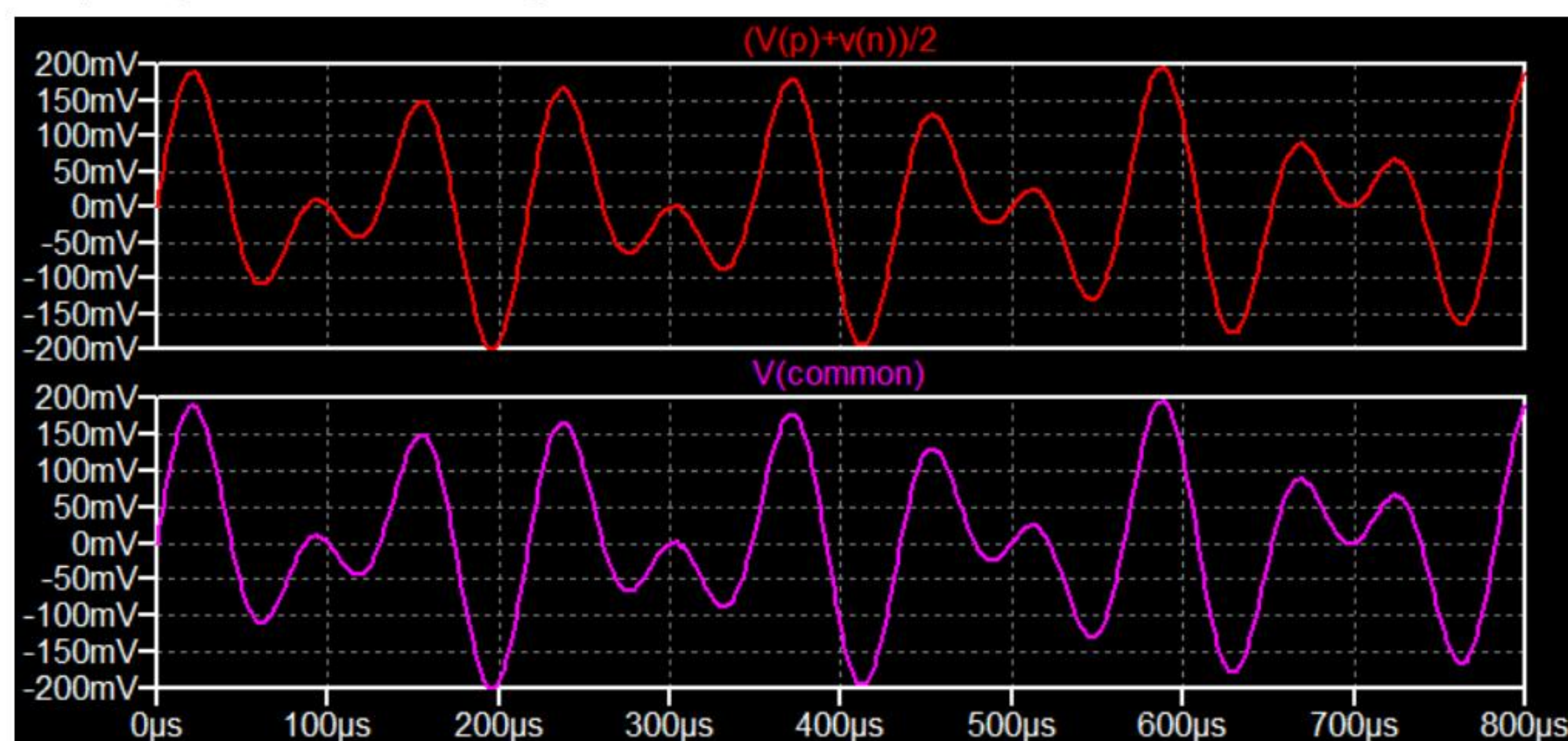


Fig.13. calculated and directly plotted common mode signal for the circuit in Fig.11.

inductively. If the wires are close together and of the same physical structure and dimensions (as is often the case), they will experience the same interference source and the same amount of coupling.

With other things being equal, the amount of voltage shift (δ) will depend on the impedance of the conductors and the circuits they connect to. If the impedances are equal, the voltage shifts on both conductors will be equal.

This is why balanced connections are important in professional audio – it is impedance balance that facilitates the removal of interference by the differential input, not the use of a symmetrical differential signal.

Differential signals provide additional advantages, so they are used in a wide range of electronic systems. Fully differential analog signal processing, where the whole circuit is differential, is often used in integrated circuits, where the circuits would typically be symmetrical between the signal paths, so inherently balanced.

In this case, some non-idealities in the circuits will also be symmetrical and thus cancel, leading to improvements such as reduced distortion.

Differential signals provide twice the signal amplitude too, which is useful in low-voltage circuits (common with modern ICs) for both analog and digital circuitry, increasing the dynamic range and improving the signal-to-noise ratio (SNR). For very high-speed digital circuitry, LVDS provides improvements in speed, timing accuracy and error reduction.

Ground noise

Another source of signal disruption in systems is ground noise, where there are fluctuations in the ground voltage difference between two systems. Ground noise appears as a common-mode signal to a differential amplifier, so is cancelled where balanced and differential connections are used. For single-ended connections, the noise is added to the wanted signal.

A common cause of ground noise in audio is ground/Earth loops where there are two (or more) ground connections between equipment, which creates a loop. This may be due to multiple cables connected (with ground connections) and/or the mains wiring in cases where the equipment is Earthed. This effect can disrupt other systems too, such as sensor readings.

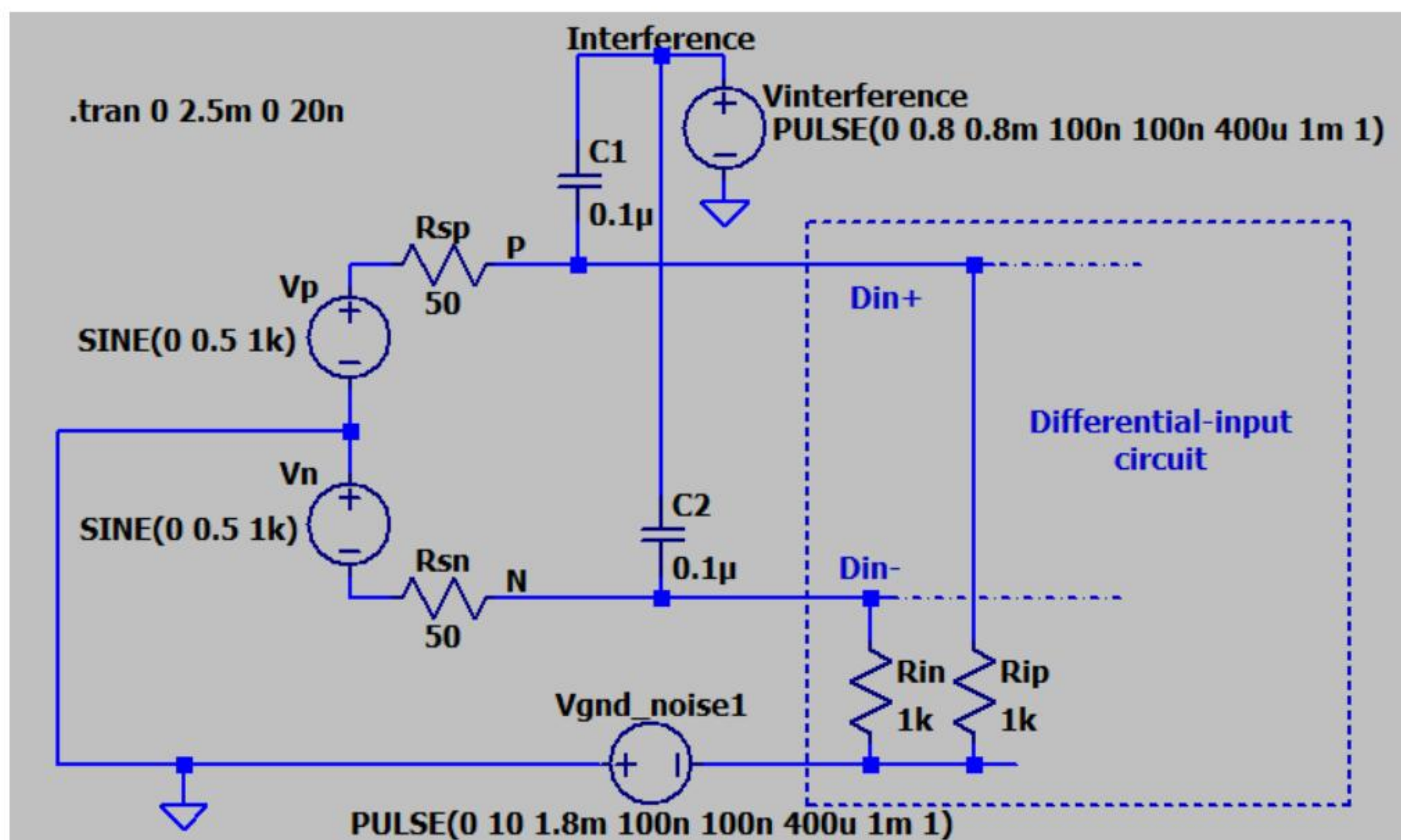


Fig.14: an LTSpice example circuit for a differential connection with source and load impedances and disturbance sources.

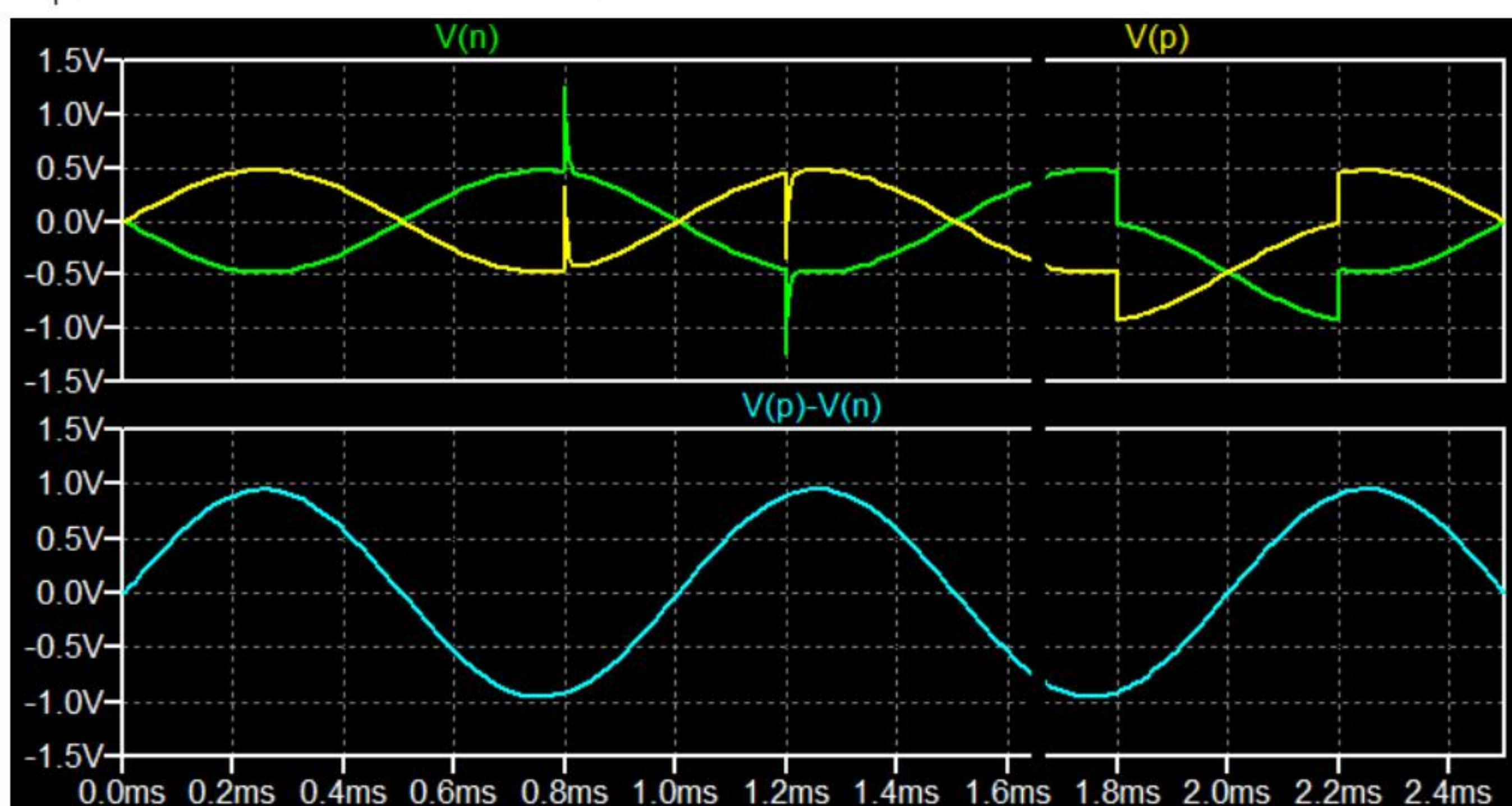


Fig.15: the simulation results from the circuit in Fig.14 with the lines balanced. The gap is where the plot switches between the interference and ground noise results.

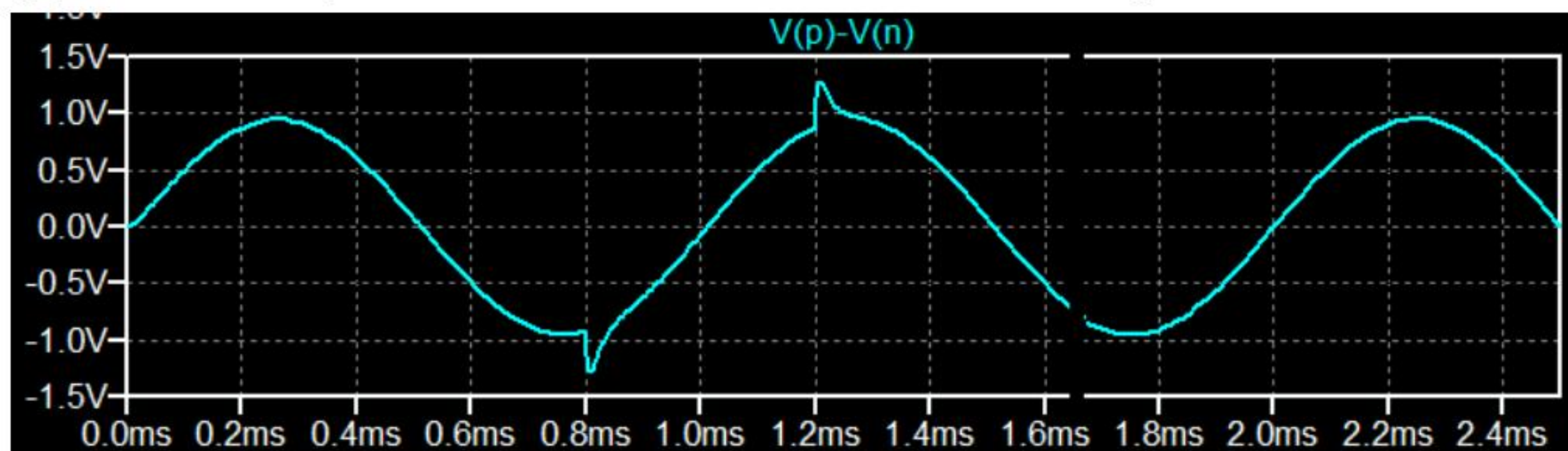


Fig.16: the simulation results from the circuit in Fig.14 with the lines unbalanced (showing interference and then ground noise).

The ground loop acts like the secondary of a transformer, with electromagnetic fields inducing currents into the ground conductors, creating a voltage across the wiring resistance. This is often at the mains frequency, so it produces hum in the audio.

Another source of ground noise is the supply current taken by the system or other systems sharing the same power wiring. As the current changes (eg, changing with the mains voltage cycle), the voltage drop across the ground resistance changes, producing a voltage shift. This can occur in a wide range of scenarios, including within ICs. Most of these effects can be avoided or mitigated through the use of balanced and/or differential connections.

LTSpice examples

Fig.14 shows a differential signal source (V_p and V_n), which is similar to Fig.11, but without the common-mode source. There are individual source resistances of 50Ω (R_{sn} and R_{sp}) before the conductors (P and N) connecting the signals to a differential input circuit that presents individual input resistances (R_{in} and R_{ip}) of $1k\Omega$ to the signals on P and N .

The fact that $R_{sp}=R_{sn}$ and $R_{ip}=R_{in}$ means that the connection from the sources to D_{in+} and D_{in-} is balanced.

As previously indicated, D_{in+} and D_{in-} are just graphical items on the schematic representing a differential amplifier input; the actual signals are P and N . The differential amplifier behaviour

Simulation files

Most *Circuit Surgery* columns feature the use of the free circuit simulation software LTSpice. It is used to support descriptions and analysis in *Circuit Surgery*.

The examples and files for this issue are available for download: <https://pemag.au/link/acb7>

will be obtained via waveform maths, $V(p)-V(n)$, as in the previous examples.

The circuit also contains two disruption sources. The voltage source **Vinterference** is coupled to P and N via capacitors $C1$ and $C2$, which represents the potential effects of an interfering electric field on the connection.

There is also a voltage source (**Vgnd_noise**) between the grounds of the source and the differential input circuit. This represents any mechanism that introduces disturbance in the ground line, producing a voltage difference between the driver and receiver grounds.

Each disruption source produces a $0.4ms$ square pulse, with the interference starting after $0.8ms$ and the ground noise after $1.8ms$. The component values and source configurations were chosen to produce clear illustrative waveforms and are not aimed at representing a specific physical circuit or conditions.

In the following simulations, only one disruption effect is simulated at a time. Either $C1$ and $C2$ are temporarily removed, with **Vgnd_noise** active, or **Vgnd_noise** is set to $0V$. Both are shown active in Fig.14 for convenience.

Fig.15 shows the results from two simulations of the circuit in Fig.14 combined on one plot. Up to the gap just after $1.6ms$, the results are for the capacitively coupled interference, and after that for the ground noise. In both cases, we can clearly see the disruption to the individual signals (upper pane, green and yellow traces) and that the differential signal (lower pane, cyan trace) is free from disruption.

Fig.16 shows another run of the same simulation, except that the resistance values in the N connection have been changed to $R_{sn}=200\Omega$ and $R_{in}=4k\Omega$. These values maintain the source and input impedance potential divider ratio, and hence the signal amplitude, but unbalance the line impedance.

In this case, we see that although the wanted differential signal on P and N is perfectly symmetrical, the interference affects the output signal because the response of the lines to the interference is not equal. The ground noise is a common-mode signal in this setup, so it is still rejected.

Fig.17 shows a single-ended, unbalanced version of the circuit in Fig.14. There is a single-ended source connected to the **Din+** differential input and the **Din-** is connected to the source ground. The signal source voltage is double that of the two individual sources in Fig.14 so that the signal is the same amplitude (at 1V peak).

The results in Fig.18 show that both the interference and ground noise disrupt the signal. Fig.19 shows a version of Fig.14 with a balanced line driven by a single-ended source (providing the same overall amplitude of 1V). Fig.20 shows the results, where we can see there is no disruption to the output. This confirms that the signal on a balanced line does not have to be differential to obtain noise immunity.

These simulations have implemented the differential input using waveform maths to provide a perfect differencing operation. Of course, real systems will not respond to their differential inputs perfectly – they will also respond to some extent to any common-mode signals, rather than completely rejecting them.

Performance in this respect is characterised by common-mode rejection ratio (CMRR) – the ratio of differential gain to common-mode gain. Given an imperfect (finite) CMRR, the common-mode signal should be reduced as much as possible to maximise the received signal quality. This is largely up to the quality of the differential amplifier in the receiver, but can be helped by good wiring practices, cable shielding and RF suppression to reduce common-mode noise and interference.

Furthermore, a good differential-output amplifier at the source will reduce common-mode signals on the line. Devices like the THAT1646 can help fulfil this requirement while providing other advantages of differential signals, such as higher signal amplitude.

THAT1646 simulation

Fig.21 shows an LTspice simulation for a minimal THAT1646 circuit. The **SENS+** inputs can be directly connected to their respective outputs, but using the capacitors (**C1** and **C2**) reduces the common-mode output (per the data sheet).

The 100kΩ load resistors (**R1** and **R2**) and input buffer amplifier (**U1**) provide more realistic I/O conditions than directly driving the device from an ideal source and leaving the outputs unloaded.

THAT Corporation's SPICE models (including the THAT1646) can be downloaded in a single zip file from <https://thatcorp.com/device-models/>

The micromodel is in the file "1646_Macro_02.lib". You also need a symbol, which can be generated by opening the .lib file in LTspice, right-clicking on the **._SUBCKT** line in the text and selecting "Create Symbol" from the context menu.

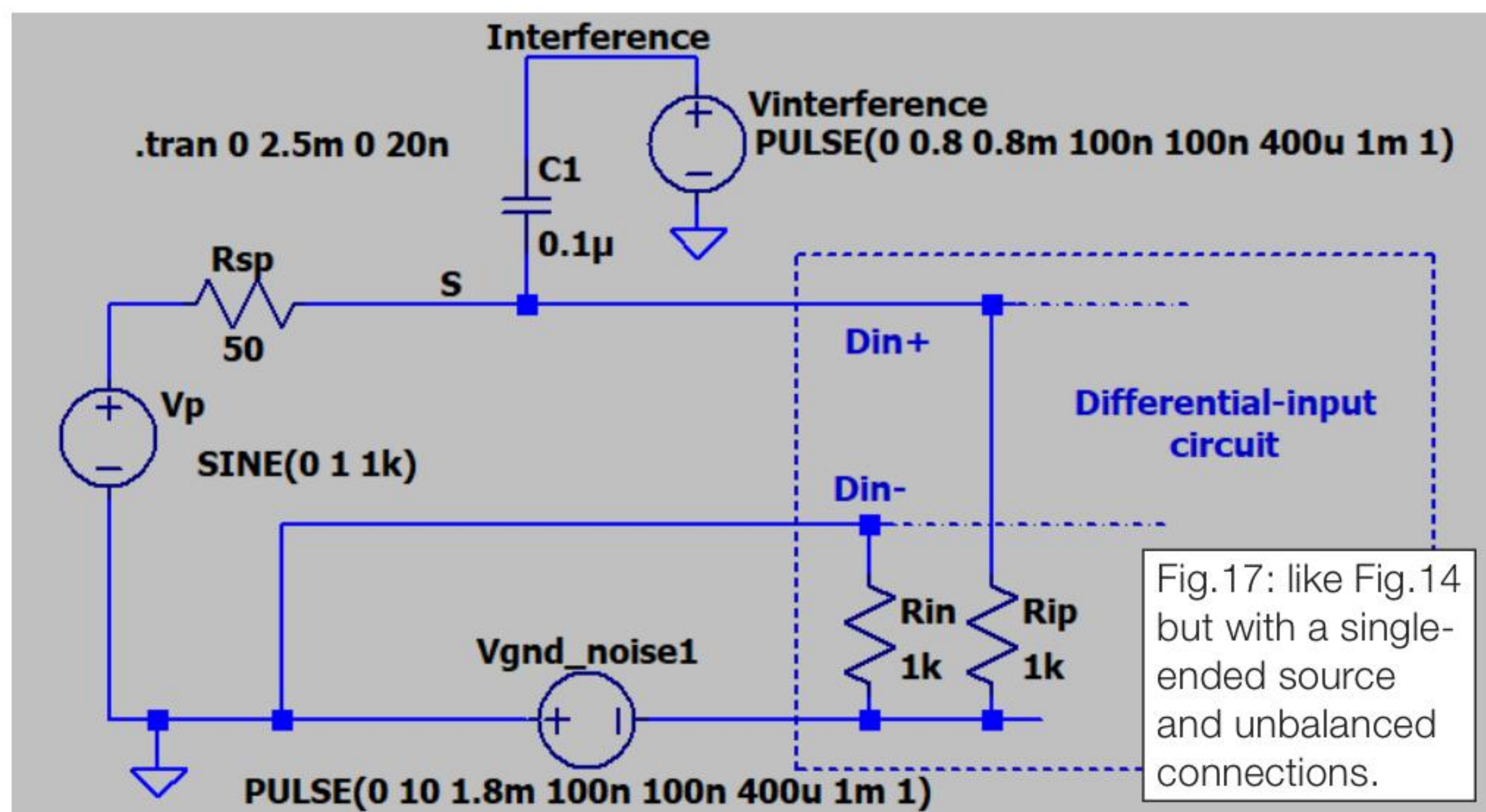


Fig.17: like Fig.14 but with a single-ended source and unbalanced connections.

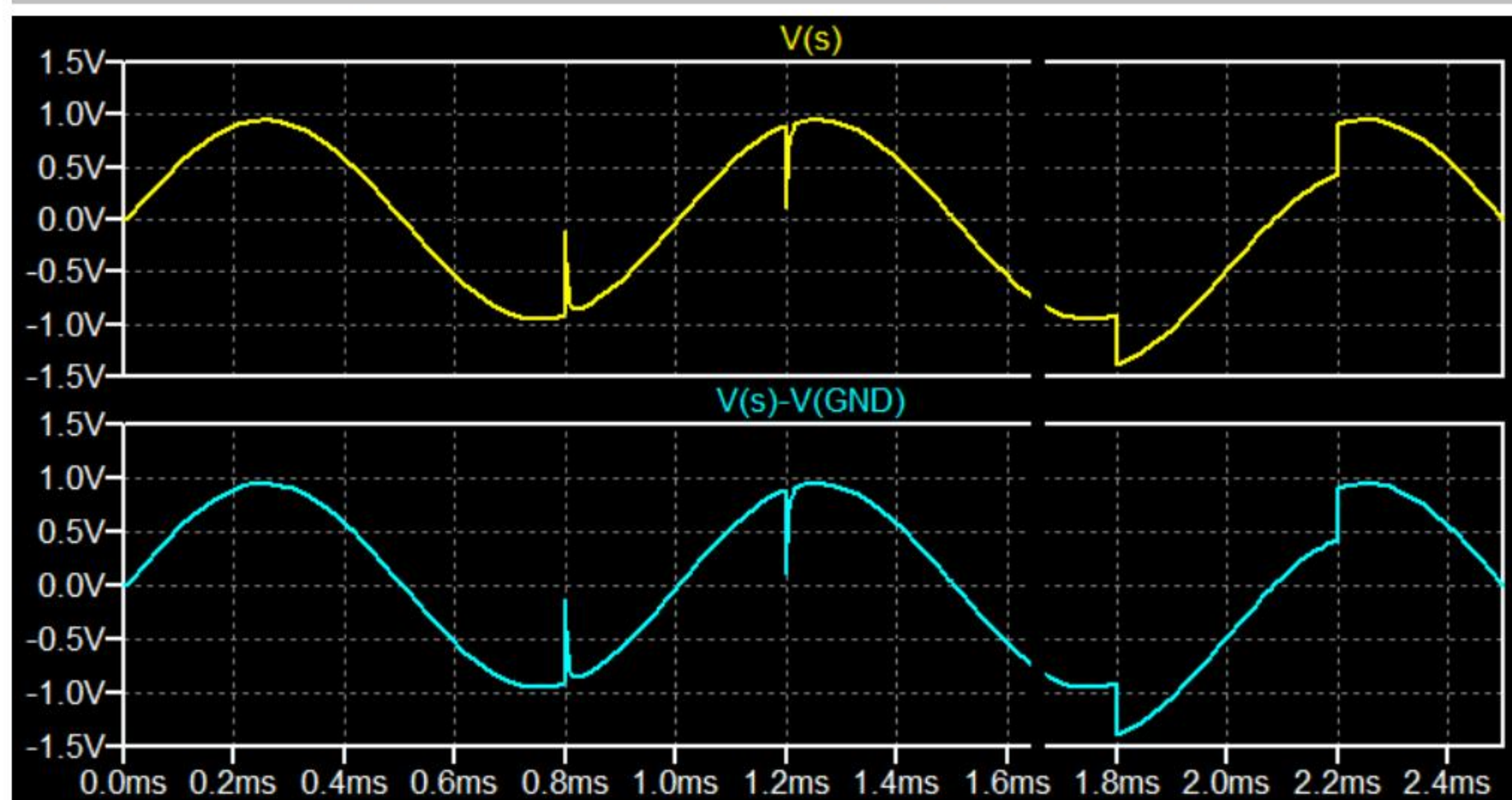


Fig.18: simulation results from the circuit in Fig.17 (interference then ground noise).

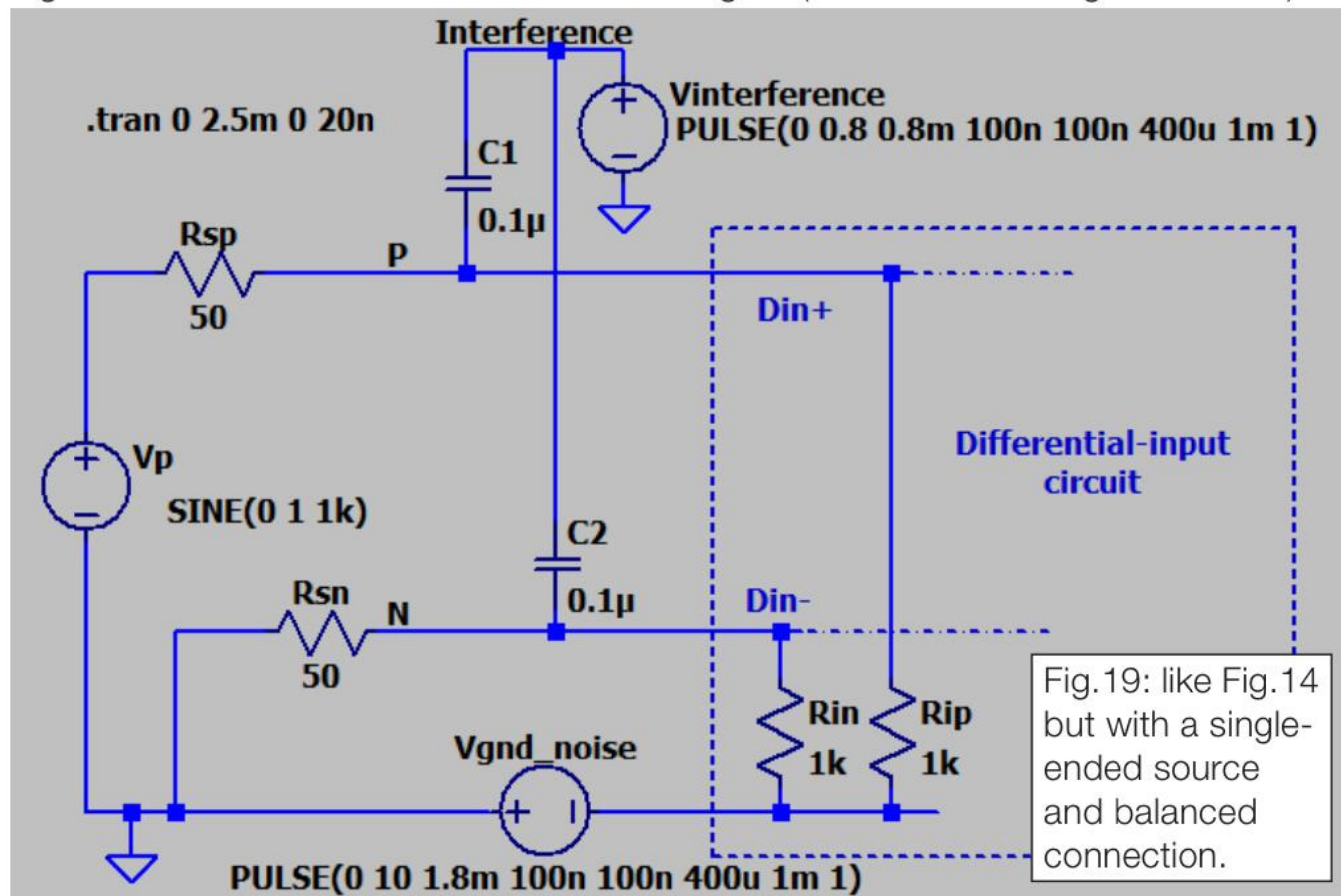


Fig.19: like Fig.14 but with a single-ended source and balanced connection.

The symbol in Fig.21 was edited to move the pins to more convenient locations.

The symbol and model files should be placed in the user library directory (see Tools → Settings → Search paths) from the main menu.

When I first ran this simulation, it was very slow, although it did complete successfully. The results in Fig.22 show the single-ended input (upper pane) and differential output (lower pane). Simulation problems can sometimes happen with the macromodel models provided

for devices. The Status Bar (see Fig.23) confirmed that LTspice was struggling.

LTspice's Status Bar

The Status Bar is a single line of text in LTspice that provides information on items at the mouse location for the currently in-focus window or gives updates on a running simulation. For schematic components, it shows editing actions and informs you of the names of nets.

After running a simulation, you also get information on operating points and

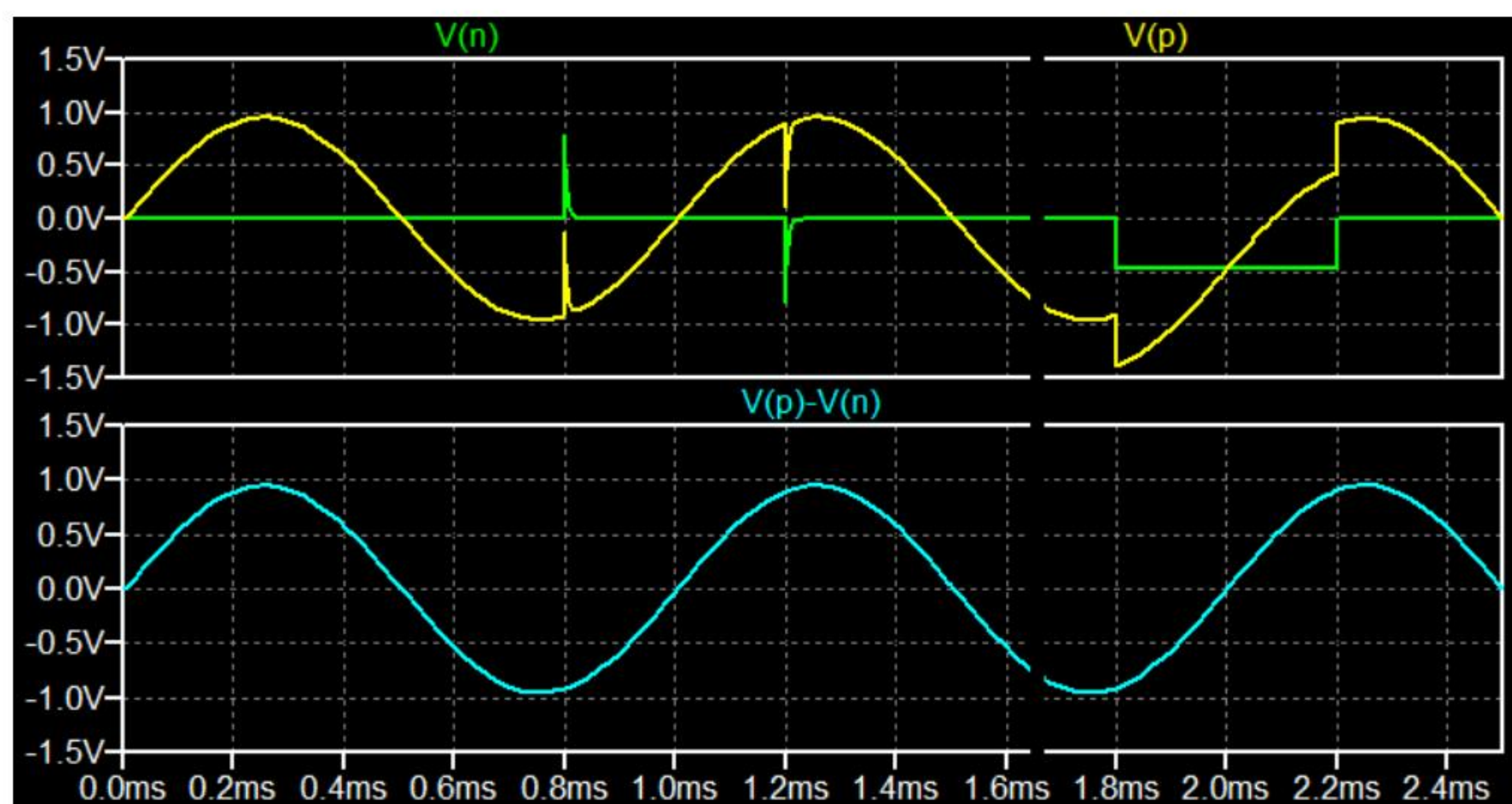


Fig.20: simulation results from the circuit in Fig.19 (interference then ground noise).

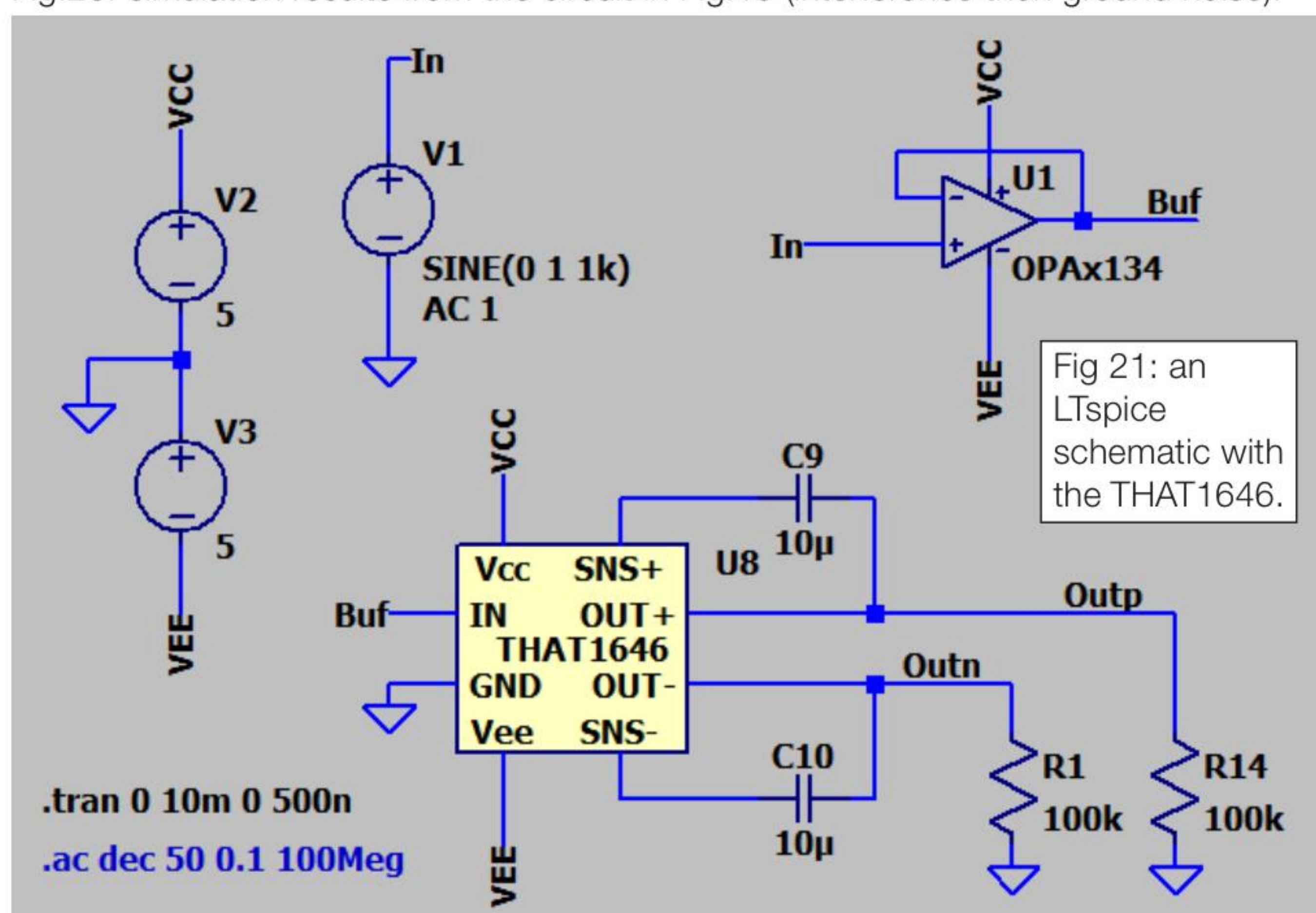


Fig 21: an LTspice schematic with the THAT1646.

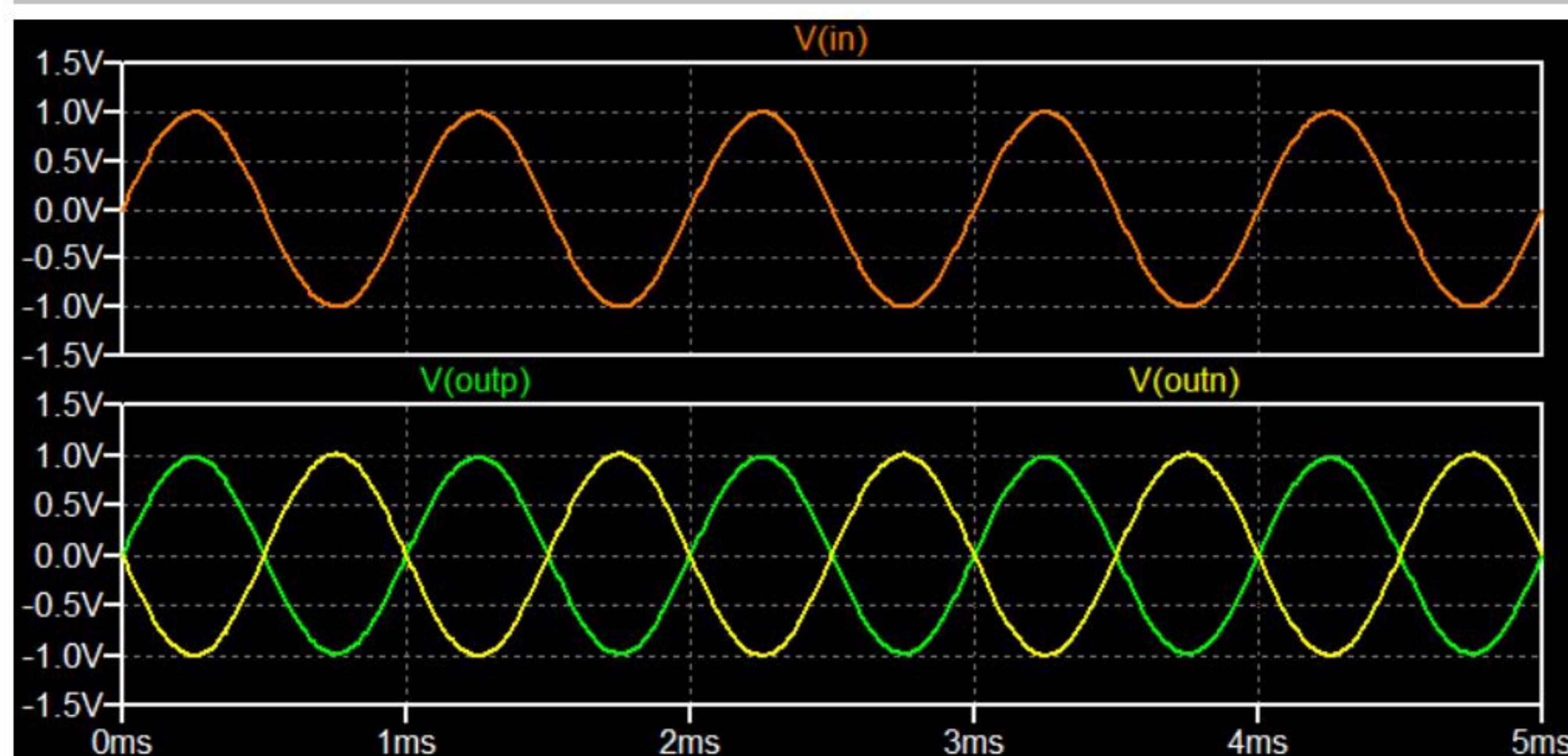


Fig.22: results from simulating the circuit in Fig.21.

Simulation Time = 499.619 µs Transient Analysis 05.0% done. Def Con 1 inter=10 fill-ins: 638

Fig.23: the LTspice status bar during a simulation.

power dissipation for components, plus prompts on actions to display results.

For waveform windows, the Status Bar shows the x and y values of the mouse position in the graph area. This is not the same as the attached cursors, as the values displayed are not locked to a trace on the graph, but this can be useful for a quick measurement by placing the mouse cursor on the relevant point on the trace.

For a simulation that takes some time

to run, you can watch the status as it progresses. Sometimes LTspice will struggle to simulate a circuit and run very slowly.

At such times, you may notice some cryptic information presented along with the straightforward time and progress data. You may see values for “Def Con”, “Inter” and “Fill-ins” (see Fig.23).

While you wait for the simulation, you might decide to pass the time by looking these up in the LTspice help, but you will

be disappointed as they are not documented there. There is very little information published on these parameters, although on a basic level they relate to how challenging LTspice is finding the simulation.

You might recognise DEFCON as the defence readiness condition used by the US armed forces, which ranges from 1 (normal) to 5 (maximum, nuclear war imminent or already started). For LTspice, Def Con is a little less dramatic – it indicates the level of difficulty the simulator is experiencing from zero upwards (it is not limited to five values).

Fill-ins may relate to the “number of elements of the system matrix that changed from zero to non-zero during LU factorization” according to a recent academic paper by Marek Nałecz investigating the efficiency of modelling techniques running in LTspice, which used the number of fill-ins as a measure of circuit complexity.

LU means lower-upper and refers to lower and upper triangular matrices. A matrix is a rectangular array of numbers, or other mathematical values, and is triangular for a square matrix if all the values above or below the main diagonal are zero. Matrix mathematics is used by LTspice to solve the multiple interrelated circuit equations to run simulations.

As an LTspice user, it is not necessary to know the details of the algorithms being used, but having some idea of the processes involved can be helpful when things go wrong. The calculations behind SPICE circuit simulation fall into a branch of mathematics called numerical methods, which provide a way of solving problems that cannot be solved analytically.

Put crudely, this involves guessing the answer and using the results to find a better approximation of the solution. The process is repeated (iterated) until the approximation is sufficiently accurate. This is not a new idea; its basis goes back to Isaac Newton (1642-1727).

Unfortunately, numerical methods can sometimes fail to find a solution or take a long time. LTspice has a variety of settings that influence its algorithms, found in Tools → Settings → SPICE. Using incorrect values here can mess things up (but there is a restore defaults button).

A useful thing to try with slow simulations is to change the Solver from Normal to Alternative (in the Engine section of the dialog). In this case, it sped up the simulation by a factor of about 25.

Conclusion

We introduced the THAT1646 IC this month, but after a detailed look at differential and balanced signals, and discussion of the status bar and slow simulation, we did not look at it in much detail. So we will continue with this topic next month.

PE

SSB Shortwave Receiver

Part 2 by Charles Kosina, VK3BAR



Introduced last month, this new Shortwave Receiver covers the entire shortwave band from 3MHz to 30MHz. It is digitally tuned and has a host of useful features like squelch, USB/LSB support, good sensitivity (a -107dBm signal gives 13dB SNR), fast or slow AGC, an RSSI display and it runs from 12V DC. This month, we describe how to build, test and align it.

This is not an overly difficult device to build, as it uses no tiny components or fine-pitch ICs. However, it has two boards that are fairly packed with SMDs plus quite a few through-hole components, so you should ideally have decent soldering skills if you're going to attempt it.

You also need some test equipment to calibrate the Radio. That includes an accurate frequency counter up to 100MHz and a signal generator that will work up to at least 30MHz that can produce a signal down to 10µV or less (or an attenuator to allow that). An oscilloscope with 100MHz or more bandwidth is also nice to have, but not absolutely necessary.

Some sheet metalwork is needed.

I recommend having a stepped drill bit or two (eg, 3-12mm & 12-24mm) on hand. A drill press would be ideal, but you can do it with a hand drill if necessary.

There are many components overall, but the values are marked on the circuit boards to ease construction. It pays to be careful as you go through the assembly process and make sure each part goes where it's supposed to. Mixing up two visually identical capacitors could be enough to prevent the radio from working.

Construction

Virtually all the components mount on two PCBs, the Control Board (Fig.14, 152.5 × 81.5mm) and the RF board

(Fig.15, 152.5 × 51mm). There are some through-hole components used, but the vast majority are SMDs, mostly passives (resistors and capacitors) in M2012/0805 packages, which measure 2.0 × 1.2mm.

These passives are on the small side if you are used to through-hole components, but we still consider them to be in the 'easier to handle' category compared to really small parts. So as long as you have the right tools, a decent amount of light and reasonable vision (or magnification), you should not find the assembly too difficult.

Similarly, the ICs are not in really tiny packages; they are mostly SOIC types with 1.27mm lead pitch, ie, half that of a through-hole chip. Again,

SSB Shortwave Receiver, part two

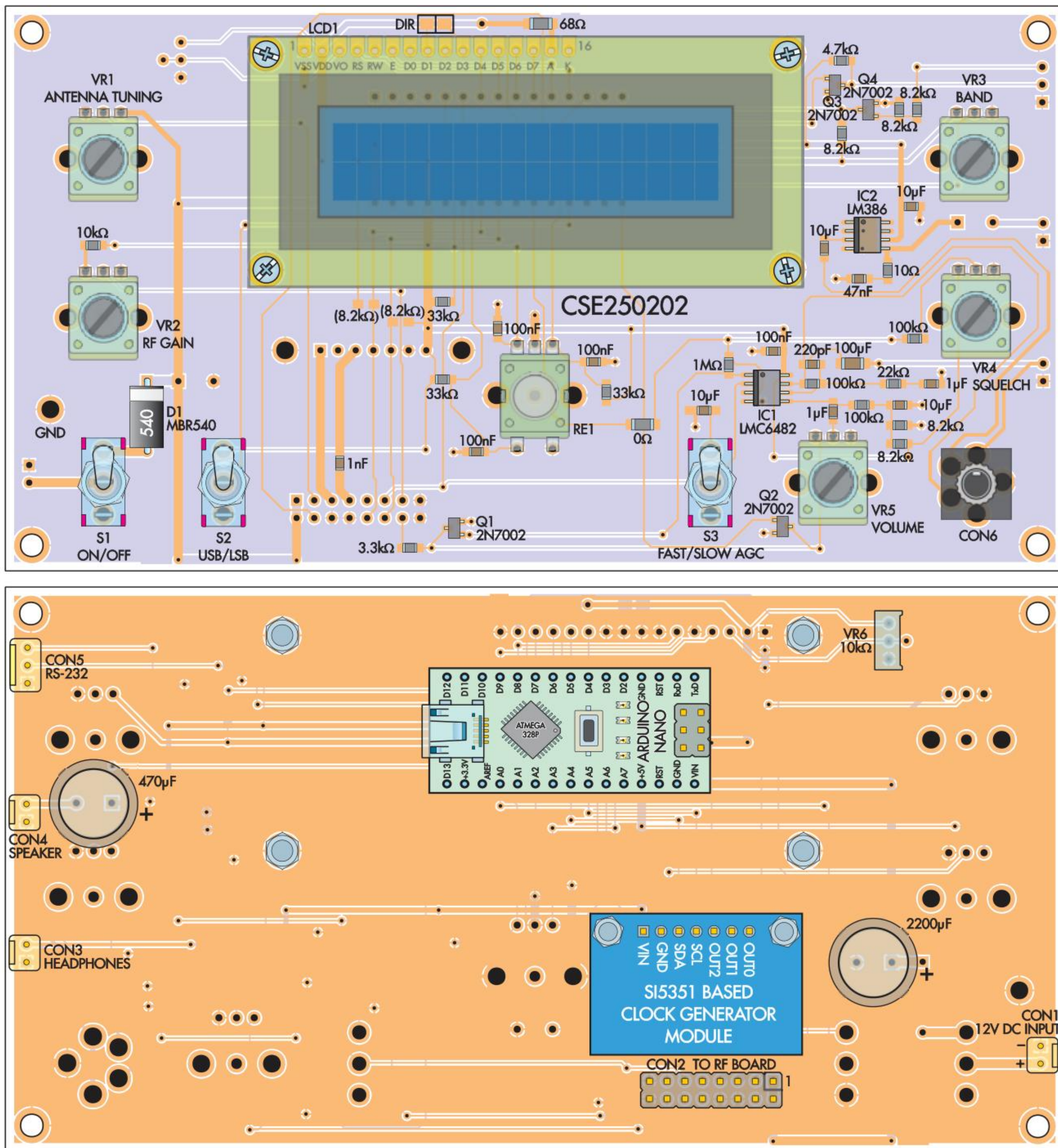


Fig.14: the Control Board has parts on both sides; fit all the SMDs first, then the through-hole parts on the underside. The ICs, diodes, electrolytic capacitors and the Arduino Nano module must all be installed with the polarities shown here for the Radio to work.

these are not what we would consider difficult-to-solder parts.

Control board

I recommend building the Control Board first and testing it before you move on to the RF Board. There are components on both sides of the board, but most of the parts, including all the SMDs, are on the front. Start by soldering the two ICs first, making sure their orientations are correct.

In each case, find the pin 1 marker (a dot or divot on the top, or a chamfered edge on the side) and make sure it's aligned as shown in Fig.14 and on the PCB silkscreen. It's possible to solder the pins of these SOIC package devices individually with a fine-tip soldering

iron. Add plenty of flux paste to make soldering easier and reduce the possibility of bridging pins with solder.

If that happens, use copper braid with a bit of extra flux paste to remove the excess solder. In fact, we usually don't bother trying to avoid bridges as it's so easy to clean them up later; we're more focused on making sure the solder flows onto each pin and pad, to avoid high-resistance connections that can be difficult to find later.

Follow with the passives using a similar technique. The resistors will be marked with codes indicating their values (eg, 10kΩ = 103 or 1002) while the capacitors will not have any markings. In both cases, it's best to unpack a single value, then fit them all as shown

on the overlay diagram so you can't get them mixed up. None of the passives are polarised so they can be soldered either way around.

Note that a few of these parts, like the 68Ω resistor and 100μF capacitor, are slightly larger than the others and so have larger pads to suit. Also, two of the 8.2kΩ resistors at centre left are not fitted (marked R10 & R20 on the PCB) as these are the I²C pull-ups and the Si5351 module has onboard pull-up resistors.

Follow by soldering the four identical Mosfets, which are all in three-pin SOT-23 packages. The pins are small but widely spaced, so this should not be too difficult.

Don't fit any of the through-hole

Constructional Project

components on the front side of the PCB (where the SMDs have been soldered) yet. The Arduino Nano and the Si5351 modules are on the back of the board and can be plugged into socket strips. This is important as if either failed, replacing them would be difficult.

15-pin socket strips are used for the Nano and one seven-pin strip for the Si5351. The only other parts on the back of the board are the headers, two electrolytic capacitors (which are polarised, so make sure they're fitted the right way around) and the trimpot for LCD contrast. All of those can be mounted now.

As well as the speaker connector (CON4), there is another two-pin header, CON3. This is connected in parallel with the headphone jack socket and may be wired to an RCA socket on the back panel for an external powered loudspeaker.

Now go back to the front of the PCB and fit the remaining through-hole components. It's important that the switches, potentiometers and encoder are square on to the board before soldering.

The way to ensure this is to attach the black front panel with 16mm tapped spacers to position these correctly; make sure that all controls turn easily before soldering. For a better appearance, rather than zinc-plated screws, I used black 6mm machine screws (which you can buy at Bunnings) to attach the front panel.

The jack socket is a unique part; ensure it is pressed firmly on to the board. Next, solder the 16-pin header to the LCD module.

Don't attach the LCD yet; clean the board with de-fluxing solvent and inspect all connections with a magnifier. Pay special attention to the solder joints on the socket strips for the Nano, as they are not accessible once the LCD

is mounted. Before any modules are plugged in, use an ohmmeter to ensure that there are no shorts from the 12V or 5V supply lines to ground.

Finally, attach the LCD module using 5mm spacers and 12mm machine screws and nuts. The Si5351a module is held in place by 6mm M2 or M2.5 screws with 11mm threaded spacers.

RF board assembly

If you want to take a break from assembly now, you could skip down to the "Programming the Nano" sub-heading, complete initial testing and calibration, then come back here when you get to the part where you need the RF board to be assembled.

The RF board overlay is shown in Fig.15. Parts are only fitted to one side of this board. As before, start with the ICs (two NE612s, one LMC6482 and the PCF8574) and ensure they are all orientated correctly as you solder them. All are in SOIC packages.

Then move on to transistors Q1-Q7; Q7 is a Mosfet, while the others are NPN RF transistors, but they all come in the same packages, so don't get them mixed up.

Follow with Q8-Q10, which have four pins since they are dual-gate Mosfets. In all three cases, the wider source lead goes towards lower right with the PCB orientated as shown in Fig.15. Fit diode D4 with its cathode stripe as shown, then REG2 after first applying a thin layer of flux paste to its pads. That will assist in soldering its tab properly. After that, solder the SMD resistors & capacitors, noting that the capacitors are again unpolarised, and all the SMDs are on the board.

Mount the axial inductors next; they have three different values, so make sure the right ones go in each location. They are not polarised, so you can fit them in either orientation.

Fit diodes D2 & D3 next; they are polarised, so ensure the cathode stripes both face to the right. After that, solder RLY1 in place with its pin 1 marking towards the top.

Next, fit VC1-VC3, which are polarised in a sense, because we want the adjustment screws to be connected to the ground pins in each case. So orientate them as we have shown.

For the varicap diode, VD1, you may get it in a two-lead TO-92 package like we did, or in an axial package, similar to a regular diode, which can be mounted vertically. Regardless, ensure its cathode lead goes to the pad marked K; with the axial package, this will be the end with the stripe.

Bend REG1's leads down and attach it to the board using an M3 machine screw and nut, ensuring its three leads go into their pads. Solder and then trim the leads. Don't do this before tightening the screw or you could fracture the leads. I used a 16mm-long screw to attach the tab as it makes a convenient ground point for testing later.

Next, install CON1 and CON2. That just leaves the crystal filter module, XF1.

The crystal filter comes with SMA sockets attached, and at least the input one has to be removed. As it's supplied, only the top connections are soldered; I used a hot air wand to carefully heat them and slide them off, but a soldering iron with a large tip could also be used instead. Take great care that other nearby components don't get removed as well.

Attach the filter to the PCB using four 10mm-long M2 or M2.5 screws, nuts and 5mm spacers. Solder wires to connect the input and output of the filter to the circuit board, one for signal and another for ground at each end. In theory, XF1 is not polarised, but it's a good idea to mount it like we did, with the angled capacitor on

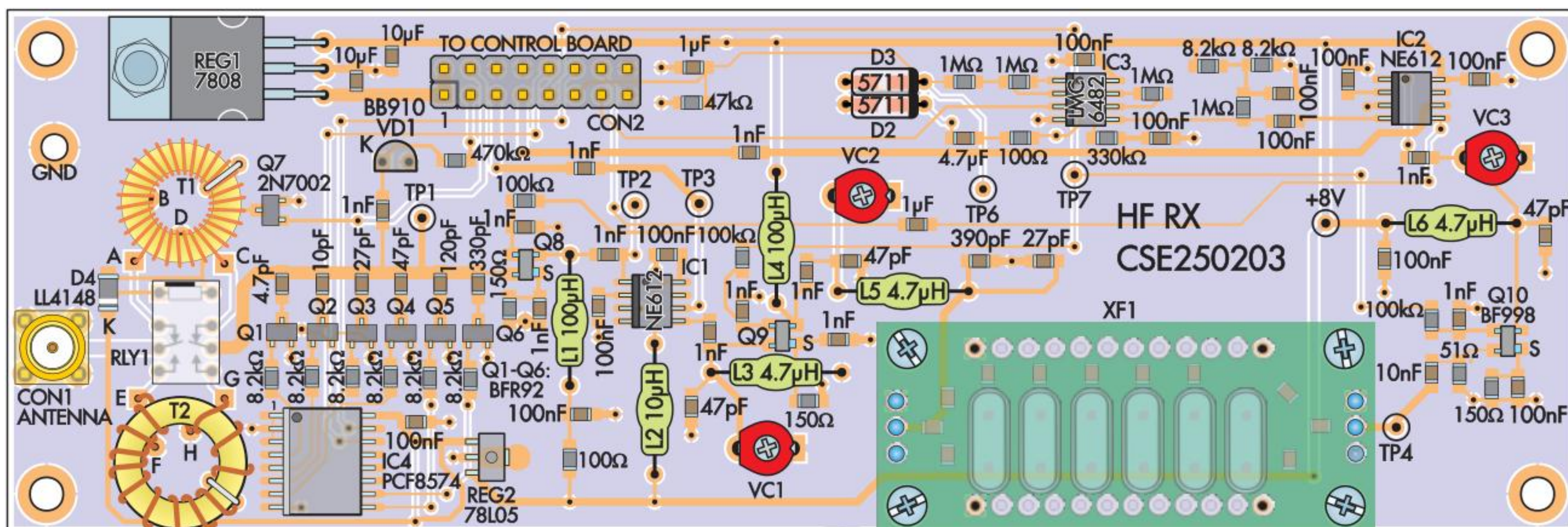


Fig.15: all the parts are on the top side of this RF Board. Polarised components to pay attention to include the ICs, dual-gate Mosfets, diodes, relay and variable capacitors. It's best to remove the SMA connectors from the crystal filter module (XF1) before mounting it on this board.

SSB Shortwave Receiver, part two

the right-hand side, since that's how we tested it.

Toroid winding

The 3-10MHz toroid (T1) needs 42 turns of 0.35mm diameter enamelled copper wire (ECW) for its secondary and four turns of the same wire for the primary. The secondary will take a little while to wind; do it first and neatly, with the turns almost touching each other. There is just enough room for that number of turns with a small gap in between the ends.

Attaching the transformers to the PCB is one of the most fiddly parts of the assembly process. T1's primary is soldered between the pads labelled A & B on the PCBs, and is wound near the 'cold' end of the secondary, while the secondary is soldered between pads C & D.

I found it easier to attach the transformers to the PCB first, solder the secondary windings, then add the primary windings. It is a bit tricky but I think it is the best approach!

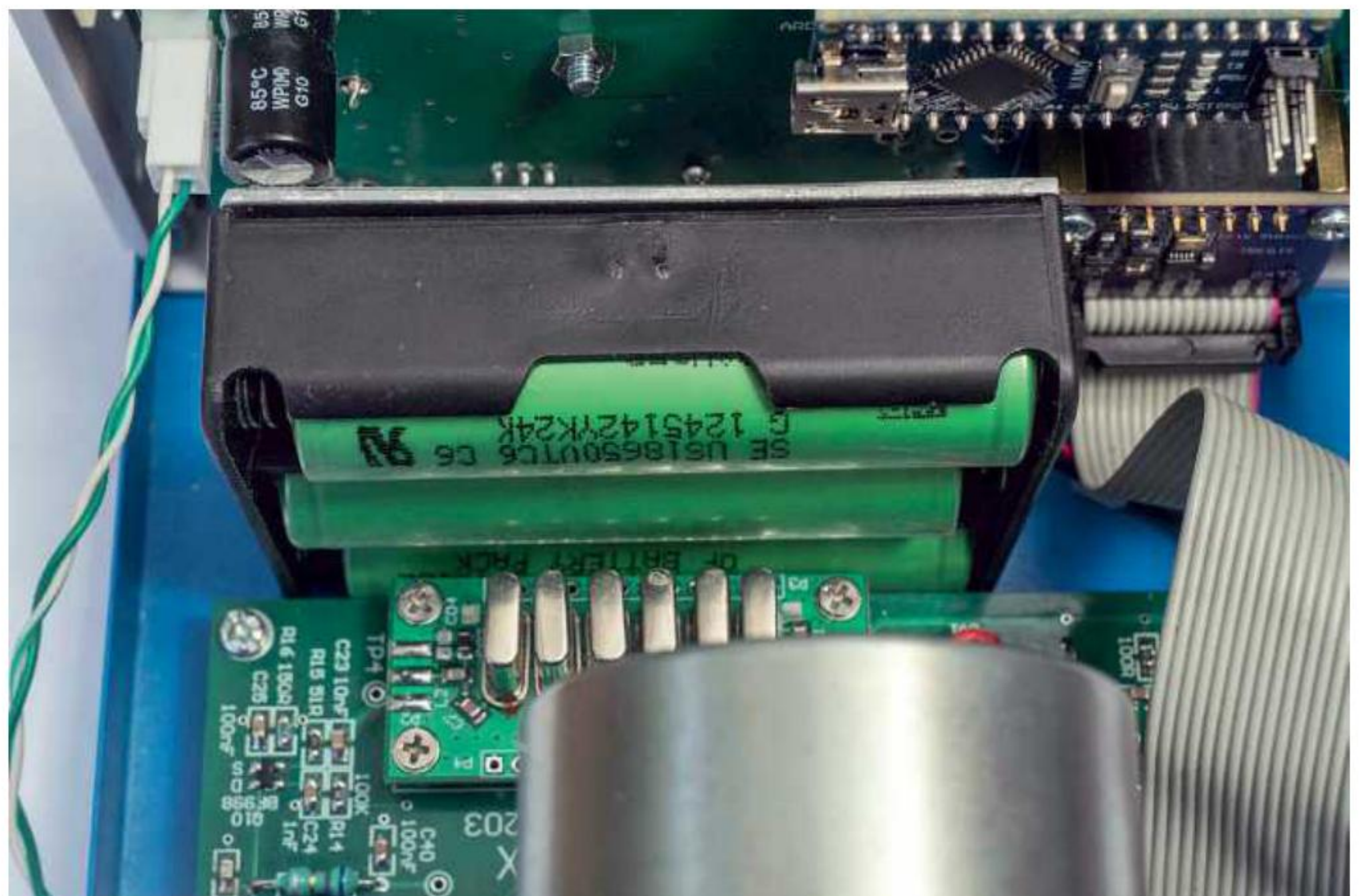
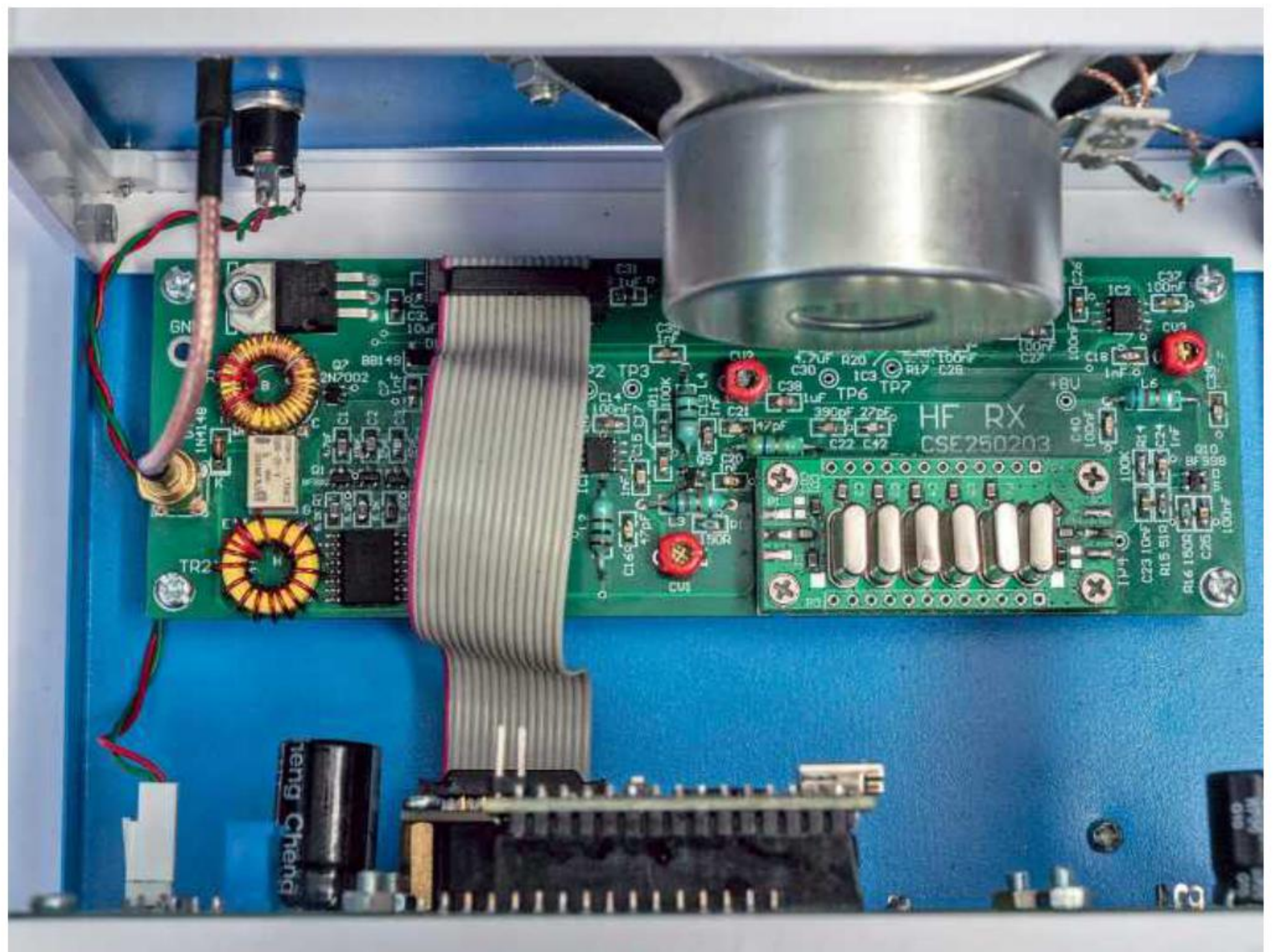
Make sure you scrape off the enamel from the wire ends before soldering them to pads A-D; otherwise, you won't get a good electrical connection and the radio won't work. It helps to tin the ECW ends after scraping them; if the solder won't stick evenly, that means you need to scrape off more enamel first.

There's also room for a tinned wire loop to help hold the toroidal core to the PCB at upper right. We recommend you add this to prevent solder joints from fracturing due to movement over time. This does not form a shorted turn as the pads it's soldered to aren't connected to anything.

The second toroid (T2) has 15 turns of 0.6mm diameter ECW for the secondary, which you should distribute evenly. Make sure the direction of winding is such that the ends go into holes G & H in the PCB correctly. Then add the two-turn primary using 0.35mm diameter ECW, scrape and tin the ends and solder it to pads E & F. Again, solder the piece of tinned wire to hold it to the PCB.

Once all components have been installed, give the board a thorough clean to get rid of surplus flux. Inspect all soldered joints and check for any shorts with an ohmmeter. Make up the connecting 16-wire cable with IDC sockets.

Use a small vice or crimping tool to



Wiring up the boards is pretty straightforward with a ribbon cable connecting the control and RF boards. There's room inside the enclosure to fit a three-cell battery holder which can be used to power the Receiver.

evenly press the parts together; make sure the cable is exactly square on to the connector. Attach the loudspeaker to the two wire connector and plug this on the control board.

Programming the Nano

The Nano should be programmed before it is plugged in. You can use the free programming software called AVRDUDESS for Windows that you can download from <https://pemag.au/link/aaxh> or use the command-line version, avrdude, if you're running Linux or macOS.

Connect the Nano via a USB cable and check what COM port it appears as. Select a baud rate of either 57,600 or 115,200 depending on the version of the bootloader in your Nano. Select the programmer Arduino for bootloader using STK500v1 protocol from the drop-down list; it may be the default. Press the Detect button and it should recognise the chip.

It is important that the programming is done in exactly the following way, making sure that the settings in AVRDUDESS are correct. Otherwise, you could end up with a bricked Nano.

Constructional Project

There are two files to be loaded: the program file, “SSB RX xx.HEX”, where xx is the version, and the “SSB RX xx.EEP” file, which is a binary file loaded into the chip’s EEPROM.

There are five boxes below Options. Tick “Disable Flash Erase”. Under the Flash box, “Format” should be “Auto (writing only)”.

Locate the HEX file to be loaded into Flash by searching for it in the square to the right of the Flash window. Locate the EEP file and place in the EEPROM window. Select Raw Binary from the drop-down list.

Tick the Write circle under Flash and press the Go box. This will result in progress messages in the bottom window.

Tick the Write circle under the EEPROM window and press the Go box. It will also have messages in the progress window.

This completes the programming, and the Nano may be disconnected from your computer and plugged into the Control Board. Make sure its orientation is correct.

Initial testing

Connect power to the board via CON1. Make sure polarity is correct; if not, nothing will happen as there is a protection diode. The LCD backlight

should be on, but there may not be any text visible. Adjust trimpot VR6 until you see text on the screen.

With the Band potentiometer (VR3) fully anti-clockwise, the top line should have 3.600.000MHz and the bottom line USB or LSB, depending on the position of switch S2. There should be a cursor visible under one of the digits. When the shaft encoder is rotated, this number should change.

Depending on the particular encoder used, it may operate backwards. In that case, bridge the two pads marked DIR above the LCD to reverse the direction.

Switch the power off and on; the screen will show “SSB Receiver” on the top line and version number on bottom line for two seconds. Toggle the USB/LSB switch and see that it changes on the screen.

Press the switch on the shaft encoder and check that each press moves the cursor to another position under the frequency. It should allow adjustments in 10Hz, 100Hz, 1kHz, 10kHz, 100kHz and 1MHz steps.

The Band potentiometer is a convenient way to cycle through the most common amateur radio bands. It sets a frequency partway through each band starting with 3.6MHz, then 7.1MHz, 10.0MHz, 14.1MHz, 21.1MHz and

28.1MHz. 10MHz is not actually in a ham band, but it has WWV transmitting time and accurate frequency information.

Using an oscilloscope and a frequency counter, check the outputs of the Si5351 module. CLK2 is the BFO and that should read 8999.6kHz or 8996.6kHz depending on the position of the USB/LSB switch.

CLK0 should have a frequency that is the sum of the currently tuned frequency plus the BFO frequency. The accuracy of these depends entirely on the 25MHz crystal attached to the Si5351 module, so you may get slightly different values. At this point, it is advisable to perform calibration.

Calibration

To calibrate the set, you need to measure the actual frequency of the 25MHz crystal on the Si5351 module. This procedure will calibrate the short-term accuracy to within less than 5Hz. Switch off the receiver, then rotate the Band potentiometer fully clockwise.

Switch it on and the top line on the LCD will show “Calibration”; the bottom line will show the nominal crystal frequency of 25,000,000Hz. In this mode, the frequency on OUT0 is set to exactly 10MHz.

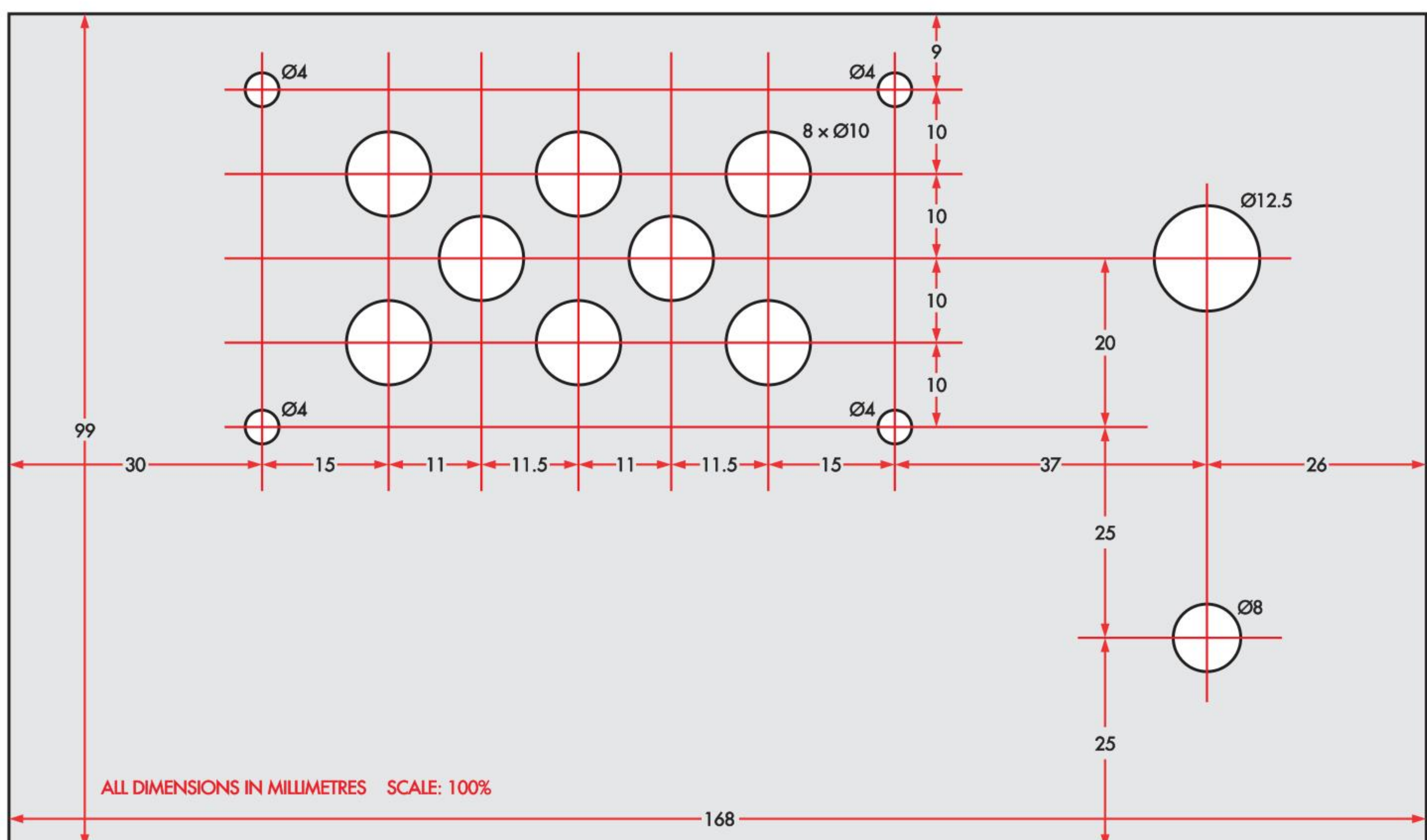


Fig.16: the recommended hole locations and sizes for the rear panel. A stepped drill bit makes drilling these straightforward. As this is at actual size, you could copy or download and print it and use it as a template.

SSB Shortwave Receiver, part two

Measure the frequency on OUT0 with an accurate frequency counter and rotate the tuning knob, which increments or decrements by 10Hz, until the output is as close as possible to 10MHz. It's possible to get to within 1Hz. It may take many turns, as the crystal could be out by more than 10kHz.

Turn the Band potentiometer anti-clockwise to leave calibration mode. The new value for the 25MHz crystal overwrites the original value in EEPROM loaded by the EEP file, and will be read every time the power is switched on.

This calibration needs to be done at regular intervals as the crystal may age and can also drift with temperature.

Case preparation

The modifications to the case involve drilling four 3mm holes in the base to attach the RF board, plus numerous holes in the back panel for the power socket, antenna connectors and the loudspeaker.

I used the Jaycar AS3025 rectangular speaker, but just about any small 8Ω speaker will be suitable. You will need to adjust the mounting hole positions if using a different speaker, though.

Fig.16 shows the suggested hole pattern. I used a stepped drill bit, as they make clean, circular holes. If you have a drill press, that would be ideal, but you can hand-drill these holes neatly if you're careful.

A word of warning! The drill can grab the plate and spin it around, possibly injuring you, so for safety, always

make sure the plate is clamped firmly while drilling the large holes.

Alignment

This should be done with the two circuit boards not yet assembled into the case to allow easier access to test points. The only adjustments on the RF board are the three variable capacitors, and they should be peaked at 9MHz. The way you do this depends on what equipment you have.

Connect the control board to the RF board with the flat cable and carefully check that the two pin 1s are connected together, ideally with the red striped side of the cable to those pins (check for continuity between the pin 1 pads on the two PCBs). If you connect the cable backwards at one end, you could do damage!

Switch on the power and set the frequency to, say, 7.1MHz or some other convenient frequency. Use an oscilloscope probe to check that you have the VFO signal at about 16MHz on TP3 and the 9MHz BFO on TP5. Connect a signal generator to CON1 with the output set to about 100μV. This is way above the lowest level, but is useful for the initial setup procedure.

Tune the signal generator for a whistle from the loudspeaker, which should be very loud. Reduce the signal generator output until you get some background hiss from the speaker. Adjust the antenna tuning potentiometer (VR1) for maximum signal, measure the DC voltage on TP6 and tweak the three trimmers (VC1-VC3) for maximum output.

That's all there is to it; the receiver should work across the whole range. You will find as you tune across 10MHz, the relay will click to switch between using T1 & T2.

Final assembly

The two boards and front panel can now be assembled into the case. The RF board uses M3 × 10mm threaded spacers to attach to the bottom of the case.

The antenna input connects via a short ready-made cable between the SMA connector to a BNC connector on the back panel. Power is from a 2.1mm or 2.5mm ID (inner diameter) DC jack on the back panel to the two pin connector, CON1, on the Control Board.

The front panel attaches to the case by four screws on the corners. Instead of using the zinc-plated screws that came with the case, I found some black screws that look better.

All the rotary controls have 6mm diameter shafts. It is preferable that these have fluted shafts, as knobs for these are more common. Still, there are a few sellers on AliExpress that have knobs with grub screws (they are listed in the parts list last month). Those were ideal for D-shaped shafts but can also be used on fluted shafts.

Conclusion

Your Radio should now be operational and you can start scanning the bands for signals! You can even use it on the go, powered from a 12V battery.

Note that while you could run this radio from a 12V vehicle battery, you must not do so (or connect it) while the engine is running as it doesn't have protection from voltage spikes. Unless you add a suitable filter, it's far safer to run it from its own internal battery. PE

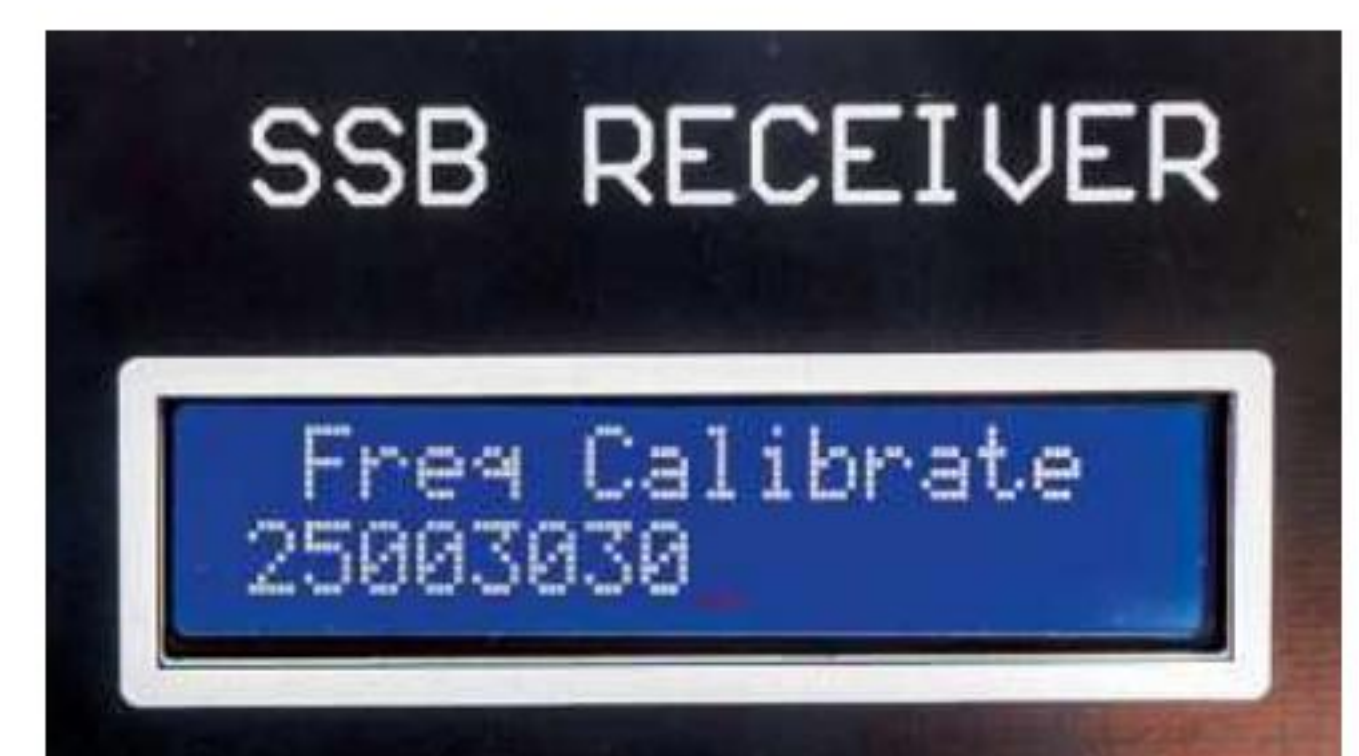


Fig.17: the front panel PCB overlay for the SSB Receiver. It is shown here at 70% of actual size.

When first running the Receiver you must calibrate it by following the text under the cross-heading "Calibration". This ensures accurate timing.

WEIRD & WONDERFUL

ARDUINO PROJECTS

Part 18: How did it get so late so soon?

I was just cogitating on the new Chronological Contraption project we briefly introduced last month. Although we are still a long way from having to concern ourselves with the container to house it, I want it to be “a thing of beauty and a joy forever,” as John Keats famously said in his poem *Endymion*. As such, ideas have started to percolate through my over-worked noggin.

Whilst ruminating on this remit, I was reminded of a previous Cunning Chronometer project that I presented in a six-sided case. I was more than happy with how this turned out, especially the back panel, which I wanted to resemble an old-fashioned radio set from the late 1920s and early 1930s. I flatter myself that I succeeded (Photo 1).

I'm particularly proud of the Morse code message in the middle, which

reads, “How did it get so late so soon?” This is a quote commonly attributed to Dr. Seuss (Theodor Seuss Geisel).

Although it's true that “beauty is only skin deep”, and it may not functionally matter whether we craft a nice case or simply plonk it in a cardboard box, I believe that it's the little details that matter when you are creating something to last. Even though few people get to see this back panel, I know it's there, and those who do really appreciate it.

That is the sort of thing I want to incorporate into the cabinet for our new Chronological Contraption.

“What was the rationale for the hexagonal interpretation of the Cunning Chronometer?” I hear you muttering to yourself. I'm glad you asked. When I originally conceived this project, it was part of a competition between a few friends, with bonus points award-

ed for originality. I initially intended to add a nine-degrees-of-freedom (9DOF) inertial measurement unit (IMU) to the little scamp.

This remains the plan, but ironically, I don't have a timeline for its realisation.

Between IMU and me...

An IMU is an electronic device that measures how something is moving and how it's orientated without requiring an external reference. At its core, an IMU measures acceleration (the rate of speeding up or slowing down) and rotation (the rate of turning).

A basic IMU typically contains a three-axis accelerometer that measures translational (linear) motion **along** the X, Y, and Z axes, plus a three-axis gyroscope that measures rotational (angular) motion **around** the same axes. This would give us what's referred to as six degrees of freedom (6DOF), because each axis of motion is independent (free) of the others.

This has nothing to do with the “six degrees of Kevin Bacon”, although we can't rule out the possibility that he's distantly related to an IMU.

Some IMUs also boast a three-axis magnetometer that measures the Earth's magnetic field, increasing the total degrees of freedom to nine, while providing some more heading information.

There are various cheap-and-cheerful IMUs on the market. Although it's a tad more expensive, my current favourite is the 9DOF Absolute Orientation IMU Fusion Breakout board from Adafruit (#4272), shown in Photo 2.

This little wonder (<https://pemag.au/link/acbo> [Amazon], [acbp](#) [Mouser], [acbq](#) DigiKey) is based on the Bosch BNO055 micro-electromechanical system (MEMS) sensor, which also includes a 32-bit microcontroller unit (MCU) for sensor fusion.

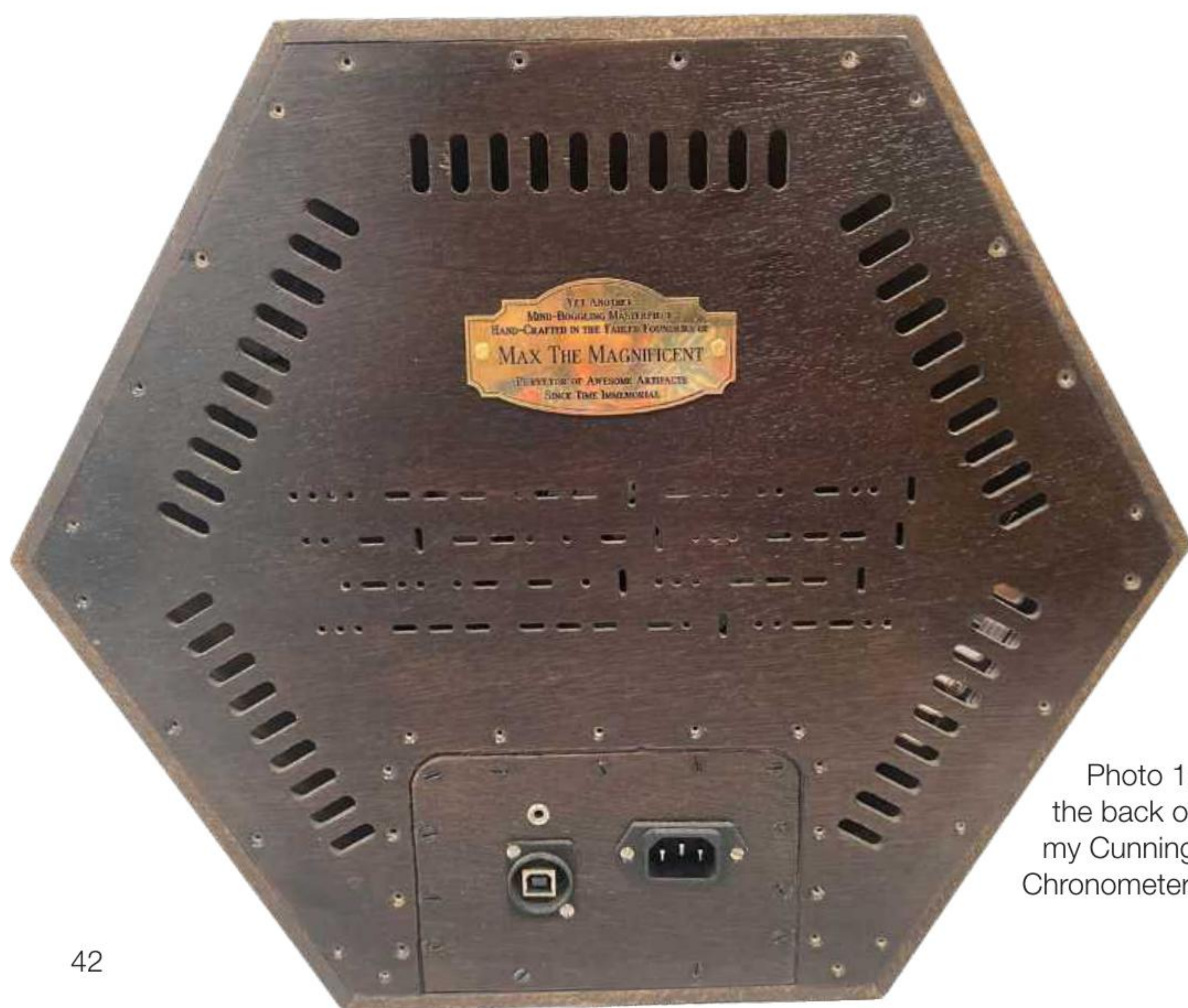


Photo 1:
the back of
my Cunning
Chronometer.

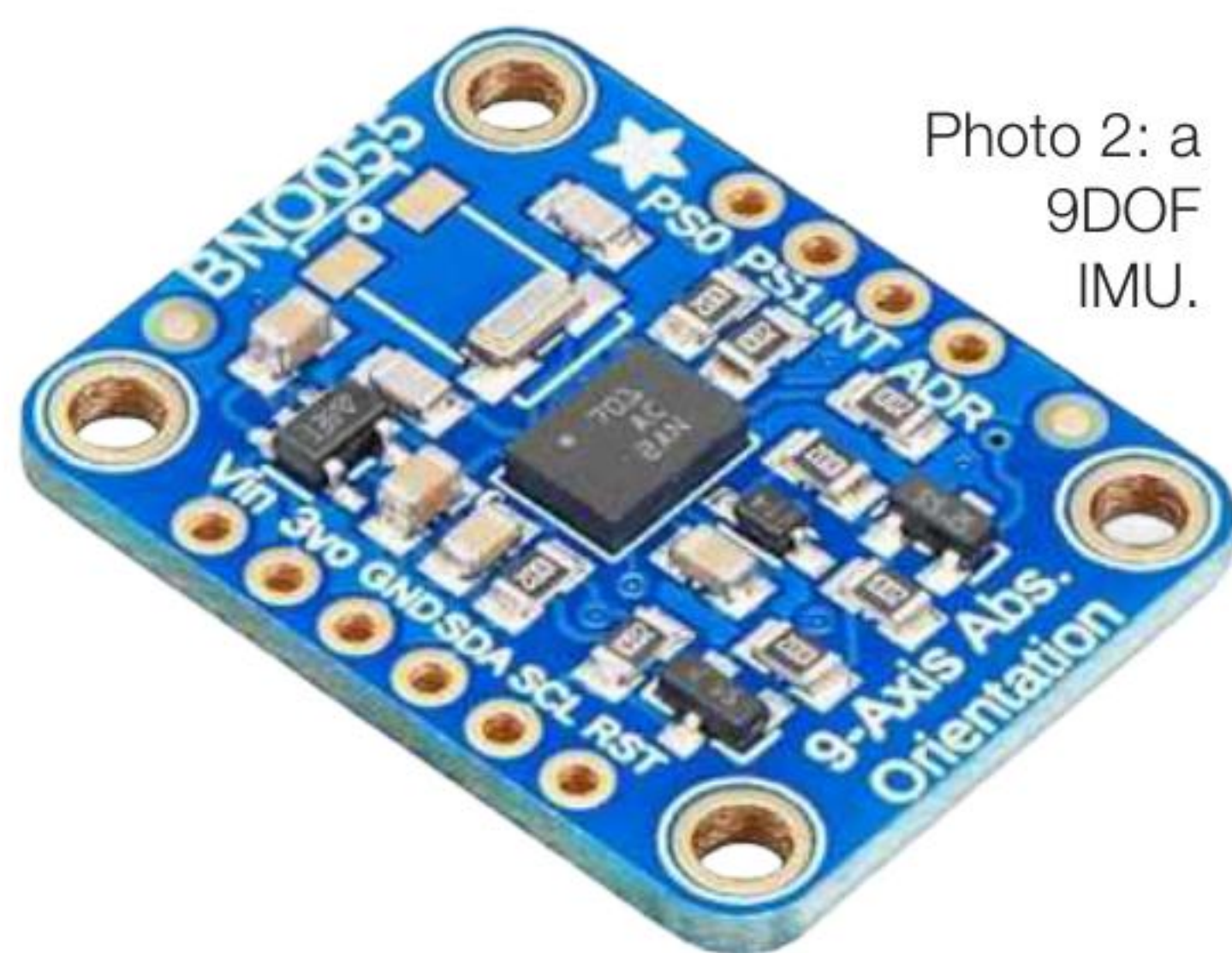


Photo 2: a 9DOF IMU.

The great thing about this Adafruit module is that they've provided Arduino code examples that get you up and running in a matter of minutes. When I eventually incorporate this little beauty into my Cunning Chronometer, I'll be able to use it to perform tricks, like rotating the case and having the display automatically counter-rotate back to its regular position.

I'll also be able to lay the clock on its back and have it turn into a compass. I'm sure that function will come in handy one day!

The reason I'm waffling about all this is that I've decided to add one of these IMUs to my retro game console. We could use it for all sorts of things, like controlling the serpent's direction in our snake game by tilting the console. We'll do that in next month's column. But first...

Let there be noise!

As you may recall from last month's column, while I was playing the latest and greatest version of our snake game, I realised that there was something missing: the tinny, quirky, 'bleepy' and 'bloopy', 8-bit Lo-Fi sounds one associates with the game consoles of yesteryear.

The more I thought about this, the more I wanted to do it, so that's what we're going to do right now. But before we leap into the fray, we need to define a few simple concepts.

'Waveform' refers to a graphical representation of how a signal changes over time. The horizontal axis shows time progressing from left to right, while the vertical axis shows the signal's amplitude (its magnitude at any point in time; for sound, this maps to loudspeaker movement). By looking at a waveform's shape, we can get a good idea of how the signal behaves and, in the case of audio, how it will sound.

Some basic waveforms include the square wave, triangle wave, sawtooth wave and sine wave, all illustrated in Fig.1. These are all examples of **periodic** waveforms, meaning that they repeat the same pattern over and over.

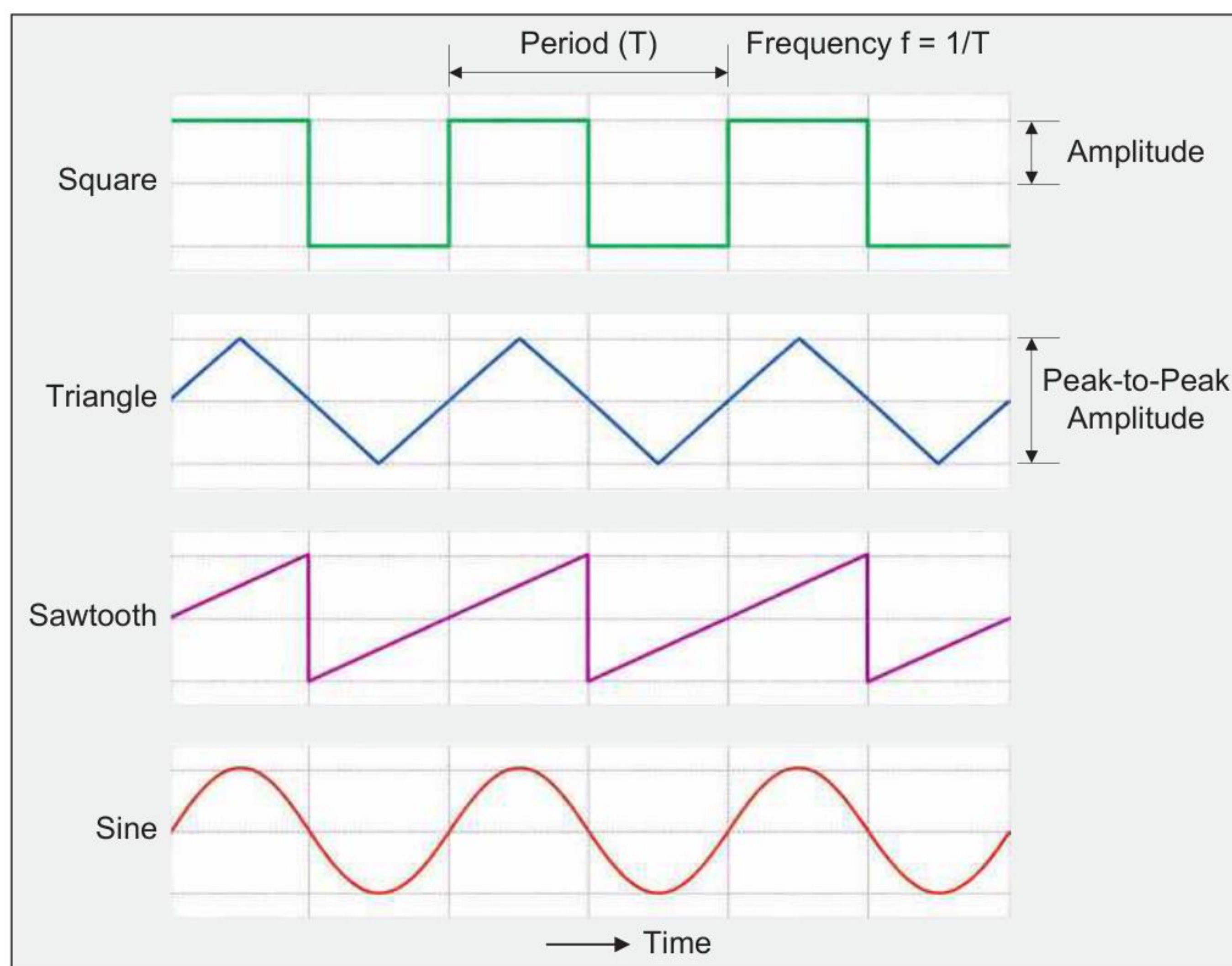


Fig.1: some fundamental waveforms.

The period (T) is the time taken for one complete cycle of the waveform, while the frequency, f, is the number of cycles that occur each second. Both are related by the delightfully simple expression $f = 1/T$, which even my poor old noggin can understand.

Frequency is measured in hertz (cycles per second). The faster the signal switches back and forth (the higher its frequency), the higher the pitch of the note we hear.

This is where things start to get interesting. The reality in plain English is as follows. A square wave is the simplest to generate, very 'digital' in character and rich in odd harmonics, making it raspy. A triangle wave is smoother, still with only odd harmonics, and has much weaker higher-frequency components (I like to think of its sound as 'soupy').

A sawtooth wave is brighter and richer, containing both odd and even harmonics; it's in between a square wave and triangle wave in sound. A sine wave is a pure tone, consisting of a single frequency with no harmonics, sounding smooth and a bit eerie (ever heard a Theremin?).

This means that in terms of perceived smoothness and refinement, we would order these waves as square → sawtooth → triangle → sine. However, if we were to consider things in terms of harmonic complexity, then the order would be sine → triangle → square → sawtooth.

Perfectly pert piezos

There are several ways to generate sound for a project like our retro game console, such as using

a small loudspeaker or a piezoelectric buzzer. A loudspeaker typically requires additional drive circuitry, whereas a piezo buzzer can be connected directly to the Arduino, so that's what we'll use.

There are two main types of piezo buzzer: active (the opinionated one) and passive (the one we like). An active piezo buzzer contains an internal oscillator circuit. These are often presented as two-pin modules: **+V** and **GND**. This is as simple as simple can be. When we apply DC power, we get a tone.

The downside is that we can typically generate only one fixed pitch (eg, 2kHz), which makes it suitable only for on/off beeping.

By comparison, a passive piezo buzzer consists of just a piezoelectric element, with no other components to speak of. These may be presented as three-pin modules: **+V**, **GND** and **IO** (the signal input). To make a sound, we need to apply a varying signal in the form of a square wave to the **IO** pin. A truly passive buzzer has just two terminals and behaves like a capacitor that makes noise.

Either way, we can generate any frequency we like (within reason), enabling us to create clicks, chirps, sweeps and other sound effects. The passive piezo buzzer's simplicity, combined with its ability to generate a variety of tones under software control, makes it an ideal choice for our purposes. It doesn't provide hi-fi reproduction, but that isn't what we want, anyway!

In fact, we already introduced such a device in June 2023 as part of our



Photo 3:
a passive
piezo-
electric
buzzer
module.

Arduino Bootcamp series (Photo 3). These little scamps are available from multiple sources, including Amazon (www.amazon.co.uk/dp/B07PRTMF89).

All we need to do is connect 0V to the **GND** pin on the left, 5V to the **VCC** pin on the right, and drive the centre pin (marked **IO**) from one of our Arduino Uno (or Nano) outputs.

Questions, questions...

The next question is: which Arduino output do we use? In the case of our game console, we can't allocate any of its digital pins because they're already tied up elsewhere, so we'll have to turn to one of the analog pins (**A0–A5**), bearing in mind that these can also serve as digital input/outputs (I/Os) if required.

To get things up and running quickly without committing myself more than necessary, I mounted my buzzer on a small board that I plugged into the 15-way connector at the bottom of my console. As you may recall, we discussed my wiring scheme for this connector in the February 2026 issue (see Fig.2).

Although the Arduino Uno boasts six analog pins, we could support only five on this connector. We decided to keep access to **A4** and **A5** in case we decide to add any I²C-based sensors later (like the IMU we discussed ear-

lier). Eventually, we decided to bid a sad farewell to **A3**, leaving us with **A0**, **A1**, **A2**, **A4**, and **A5** to play with.

But wait! We can't use **A0** because, as we discussed last month, we are already using this pin to seed (initialise) the Arduino's pseudo-random number generator (PRNG) with a suitably unpredictable value.

In my case, I decided to use **A1**. However, if I eventually decide to mount my buzzer inside the console, I would swap this to **A3**, thereby freeing up the **A1** pin on the 15-way connector.

Engineering onomatopoeia

Since we're here, let's briefly dip our toes into the onomatopoeic engineering terminology waters. Onomatopoeia refers to words that imitate sounds, like buzz, click, beep and bang.

We'll start off with three basic sound effects: a short, sharp "click" to keep time with the snake's slithering progress; a jaunty "chirp" when it snaffles a snack; and a suitably mournful "wail" when it manages to collide with something it really shouldn't, thereby meeting an untimely end.

The Arduino environment provides a handy function called **tone()** that makes generating simple sounds remarkably easy. When we call this function, we specify a pin and a frequency, and the Arduino produces a square wave on that pin at the requested frequency. This will cause our piezo buzzer to vibrate and generate a sound.

We know to our cost that changing too many things at once is problematic. So, as usual, we'll take things step by step. Let's start by generating a "click" sound in isolation. I just created a quick test program in the file named **CB-jun26-code-01.txt** (all files mentioned in this column are part of the June 2026 download package from the PE website at <https://pemag.au/link/acbm>).

This code is shown in Listing 1(a). We're using the convention

that the listing number (1 in this case) corresponds to the numerical part of the matching code file name ("01" here). Although we could have easily implemented everything in the **loop()** function, it's a better idea to split the **PlayClick()** function out to facilitate reuse in our snake program later.

We can call **tone()** in two ways. The simplest form, **tone(pin, frequency)**, as seen on line 22, starts the tone and lets it run continuously. This is the approach we're using. In this case, the sound will continue until we explicitly stop it using **noTone(pin)**, as on line 28.

Alternatively, we could specify a duration (in milliseconds) using **tone(pin, frequency, duration)**. In this case, the Arduino automatically stops the tone after the specified time, which can be convenient for short sound effects like clicks and chirps.

Because all of this is handled 'behind the scenes' using timers and interrupts, the overhead is minimal, making **tone()** an efficient way to add simple sound effects to our projects.

Run this program to see what you think. You might try varying the frequency and duration of your clicks, remembering that a typical passive piezo buzzer works best in the 1-5kHz range.

Feeling chirpy

As uplifting as it is, we're not talking about The Wiggles' *Feeling Chirpy* song from their *Top of the Tots* album and video. Rather, our next task is to emit a cheerful "chirp" sound when our snake snaffles a snack.

We can create such a sound by sweeping the frequency upward over a short period, using the code in the file named

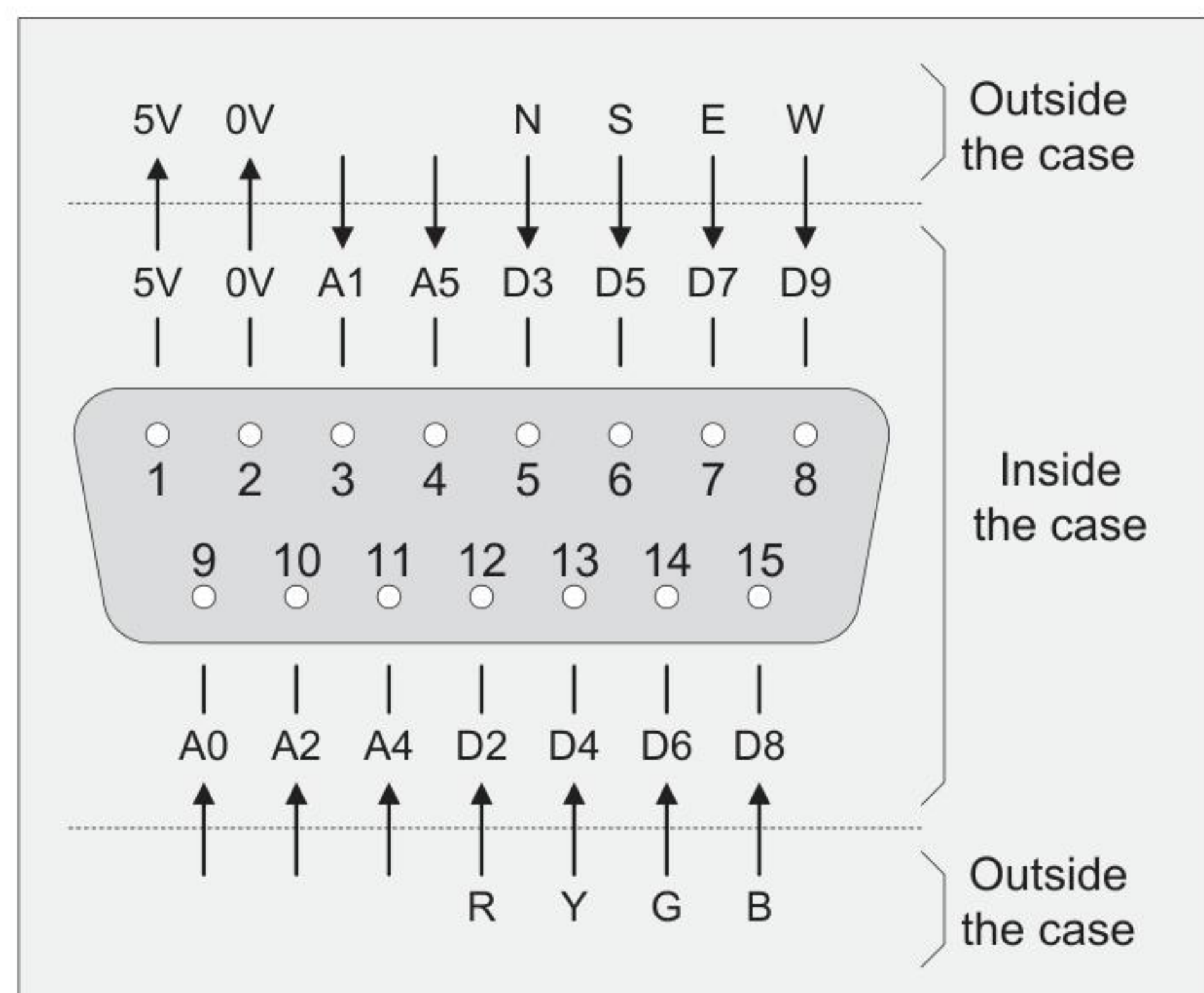


Fig.2: the wiring
on my 15-way
connector.

```
1 // Click sound test program
2
3 int buzzerPin = A1;
4
5 void setup()
6 {
7   pinMode(buzzerPin, OUTPUT);
8 }
9
10 void loop()
11 {
12   // Play click repeatedly for testing
13   PlayClick();
14
15   // Wait before repeating
16   delay(500); // Pause between clicks
17 }
18
19 void PlayClick ()
20 {
21   // Start tone at 2kHz
22   tone(buzzerPin, 2000);
23
24   // Wait a very short time
25   delay(5); // short, sharp click
26
27   // Turn tone off
28   noTone(buzzerPin);
29 }
```

Listing 1(a): our click sound test.


```

1 // Chirp sound test program
2
3 int buzzerPin = A1;
4
5 void setup()
6 {
7     pinMode(buzzerPin, OUTPUT);
8 }
9
10 void loop()
11 {
12     // Play chirp repeatedly for testing
13     PlayChirp();
14
15     // Wait before repeating
16     delay(1000); // Pause between chirps
17 }
18
19 void PlayChirp ()
20 {
21     // Sweep from low to high frequency
22     for (int freq = 1000; freq <= 3000; freq += 100)
23     {
24         tone(buzzerPin, freq);
25         delay(20); // Short delay for each step
26     }
27
28     // Turn tone off
29     noTone(buzzerPin);
30 }

```

Listing 2(a): our chirp sound test.

CB-jun26-code-02.txt. Although each individual step is just a simple tone, our ears blend them together into a single rising sound that gives a pleasing sense of success.

Consider the **PlayChirp()** function in Listing 2(a). In this case, we start at 1000Hz and climb to 3000Hz in 100Hz steps, each lasting 20ms, resulting in a short “wee-ooop!” style chirp.

Take this program for a spin, then experiment by adjusting the chirp’s character. For example, you can make it faster and snappier by decreasing the delay value to 10, or slower and more expansive by increasing it to 30.

You can generate a higher-pitched chirp by changing the loop control to **for (int freq = 1500; freq <= 4000; freq += 100)**. Alternatively, for a more cartoony effect, try using larger steps, such as **for (int freq = 1000; freq <= 3000; freq += 200)**. And for a slightly more polished version, reduce the delay to 15 and use **for (int freq = 1200; freq <= 2800; freq += 50)**.

The moment of doom

So we find ourselves at the moment when our brave little serpent comes to a sticky end. This is where we get to indulge in a “uh-oh...” sound effect. To signal the snake’s untimely demise, we simply reverse the process we used in the “chirp”, sweeping the frequency downward (see the file named **CB-jun26-code-03.txt**).

By stretching the timing slightly, we can give the sound a more dramatic, sinking quality—rather like the snake realising, just a fraction too late, that things have gone horribly wrong. Con-

sider the **PlayDoomGloom()** function in Listing 3(a).

This time, we start at a bright, attention-grabbing 3000Hz and fall to about 300Hz, which is as low as the buzzer will reasonably tolerate. We also use a slightly longer delay of 30ms, which gives the effect a more ‘sinking’ feel. The result is a classic “wahhh...” game-over sound.

Fire up this new version of our program and tickle your ears with its rueful refrain. Try making some tweaks for extra drama. For a faster panic drop, decrease the delay to 15; for a slower, more mournful effect, increase the delay to 40.

Use bigger steps to achieve a more ‘8-bit’ feel using **for (int freq = 3000; freq >= 300; freq -= 200)**. Since piezo buzzers can get a bit sulky at low frequencies, you can achieve a cleaner sound by not dropping too low, using **for (int freq = 3000; freq >= 600; freq -= 100)**.

Last but not least, if you want something a little more expressive and theatrical, you can try slowing the sound as it falls, as in the file named

CB-jun26-code-04.txt.

This requires just a small tweak to our **PlayDoomGloom()** function, as illustrated in Listing 4(a). As you’ll see (well, hear), this gives it a distinct “oh no... oh no... oh no...” quality as it tails off.

In this case, we commence with a 20ms delay between steps, increasing it by 2ms for each step. Do you wonder what it would sound like if we increased it by 3ms, 4ms,

or even 5ms each step? There’s only one way to find out!

It’s music to my ears!

I just added our sound effects functions to the latest and greatest version of our snake program, in the file named **CB-jun26-code-05.txt**. I tell you, it’s amazing how much depth a little sound adds to the gaming experience.

What’s that, you say? You’d like some homework? Well, I’m nothing if not obliging. How about you try to come up with some effects of your own? Perhaps a sound we could play when we start the game or move to a new level. How about a sound for when we beat our previous score? I’d love to hear what you come up with.

If we decide to push things further in the future, there’s a whole field of rabbit holes waiting for us, including playing little tunes, multi-note effects and non-blocking sound routines. We could even have some background music as the snake slithers around! But honestly, what we have now is

```

1 // Doom-and-Gloom sound test program
2
3 int buzzerPin = A1;
4
5 void setup()
6 {
7     pinMode(buzzerPin, OUTPUT);
8 }
9
10 void loop()
11 {
12     // Play D&G repeatedly for testing
13     PlayDoomGloom();
14
15     // Wait before repeating
16     delay(1500); // Pause between D&G
17 }
18
19 void PlayDoomGloom ()
20 {
21     // Sweep from high to low frequency
22     for (int freq = 3000; freq >= 300; freq -= 100)
23     {
24         tone(buzzerPin, freq);
25         delay(30); // Increase for dramatic effect
26     }
27
28     // Turn tone off
29     noTone(buzzerPin);
30 }

```

Listing 3(a): the doom and gloom sound test.

```

19 void PlayDoomGloom ()
20 {
21     int delayTime = 20;
22
23     // Sweep from high to low frequency
24     for (int freq = 3000; freq >= 300; freq -= 100)
25     {
26         tone(buzzerPin, freq);
27         delay(delayTime);
28         delayTime +=2; // Gradually slow down
29     }
30
31     // Turn tone off
32     noTone(buzzerPin);
33 }

```

Listing 4(a): doomier and gloomier.

exactly right for our retro game aesthetic.

Tick-tock

Let’s get back to our new Chronological Contraption project. As discussed last month, this is intended to mark the years, months, days, hours, minutes, and seconds until the start of my 100th birthday celebrations, but you can use it for anything you want (it could even act as a clock at a pinch).

We could implement this using any form of display, such as the 7-segment light-emitting diode (LED) displays we’re using on the retro game console. However, I decided to use Soviet-era vacuum fluorescent displays (VFDs), similar to the one shown in Photo 4.

A VFD is essentially a vacuum tube containing phosphor-coated segments. Different combinations of segments can be activated to form numbers. They come in all sorts of shapes, sizes and form factors (this is true of both the tubes and their segments).

In our case, we’re using the IV-11 type, which is conceptually similar to our 7-segment displays, including a decimal point segment. These IV-11 tubes are readily available as surplus NOS (new old stock) parts from hobby electronics sellers and on eBay.

There are several reasons we chose VFDs, starting with their bright, crisp and easy-to-read display, which offers excellent visibility, a wide viewing angle and fast response times. They also have that wonderful ‘retro glow’ that’s hard to describe and impossible to resist.

Compared to Nixie tubes (for which I also have a great fondness), VFDs are less expensive and operate at far lower voltages (around 20–30V instead of the 170–200V demanded by Nixies), making them a lot easier to drive. We will delve deeper into VFD technology next month (these really are awesome little devices) but we currently have other fish to fry.



Photo 4: a vacuum fluorescent display (VFD) inside a glass tube. Source: <https://display-tubes.org/vfd/reflector-iv-12/>

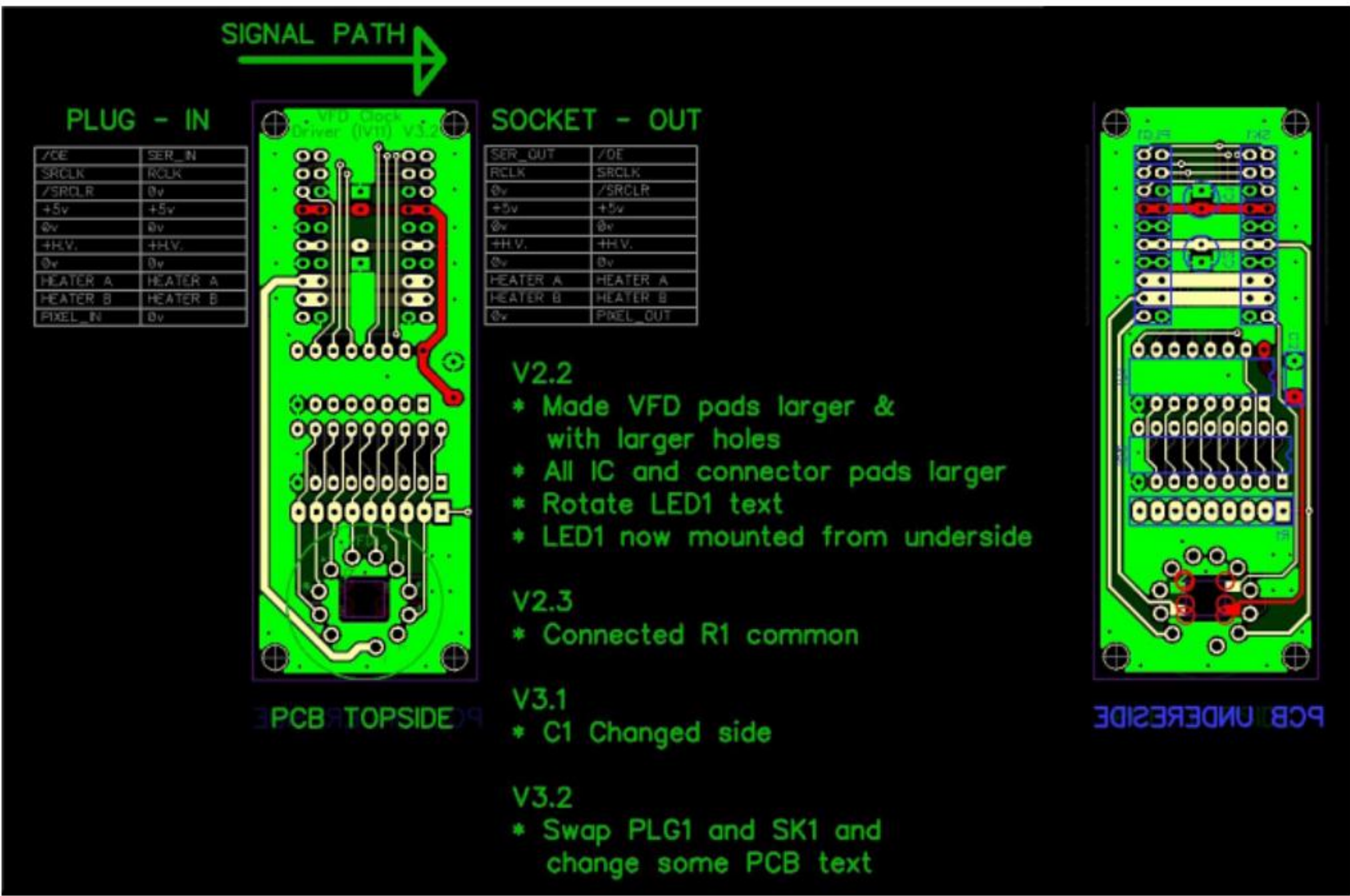


Fig.3: the layout of the main PCB; top side (left) and bottom side (right).

How many digits?

Since this is a pet project, I’ve decided to go all-out, using 14 tubes to display years, months, days, hours, minutes and seconds in a YYYYMMDD hhhmmss format.

Of course, you could always use fewer tubes. For example, you might reduce the year to a two-digit YY format, implicitly assuming the most significant digits are 20.

There are all sorts of tricks we could play here, such as using only six digits to alternate between displaying YYMMDD and hhhmmss.

Following this train of thought, we could use only two digits to switch between displaying YY, MM, DD, hh, mm and ss. Still, that may be a little confusing.

A cunning plan

As usual, my ‘partner in crime’ on this project is my friend Joe Farr. It’s great to have someone to bounce ideas off. It’s also great to have someone to blame if things go pear-shaped, although, to be honest, that’s more of a benefit to Joe than me.

To accommodate readers who want to build Chronological Contraptions of various sizes, Joe and I came up with a cunning plan worthy of Baldrick. The idea is to have a small printed circuit board (PCB) for each tube, with connectors that let them plug together to form chains of different lengths.

The topside and underside of this board, which Joe and I typically refer to as the “main” board, are illustrated in Fig.3. Don’t worry about the nitty-gritty details at the moment; we’ll look into this in more depth next month. All we need to do now is wrap our brains around a few high-level details.

Let’s start with the fact that the signal path goes from left to right. At the top

of the board, we have two 2×10 (20-way) right-angle PCB connectors. The input connector on the left is a male plug, while the output connector on the right is a female socket.

The reason we decided to do things this way (rather than go with the socket on the left and the plug on the right) is to prevent potential problems with exposed plug pins (for once, we were thinking ahead). At the bottom of the board, we find our 11-pin VFD tube (we’ll talk about what all these pins do later).

In the middle of the board, we have two integrated circuits (ICs) along with a single inline (SIL) resistor network. We opted to use through-hole (TH) components rather than their surface-mount device (SMD) counterparts to make it easy for anyone to build these. Once again, we’ll talk about all these components in more detail later.

For now, we need only note that the 16-pin IC closest to the connectors is a trusty 74HC595 serial-in, parallel-out (SIPO) shift register. We’ve used it in several projects to date. The 18-pin IC snuggled up to the tube is the ULN2803A, which acts as a buffer/driver, allowing the 5V logic signals from the ‘595 to control the higher-voltage (~25V) signals required by the VFD.

Only the VFD tube is mounted on top of the board; all other components, including the connectors and capacitors, are on the underside. Note that the image of the underside of the PCB, as shown on the right in Fig.3, represents the view as if we were looking **through** the board; hence the reversed text. We’d see a mirror-image of that if we physically flipped the board over.

Let there be LEDs!

I can almost see a frown furrowing

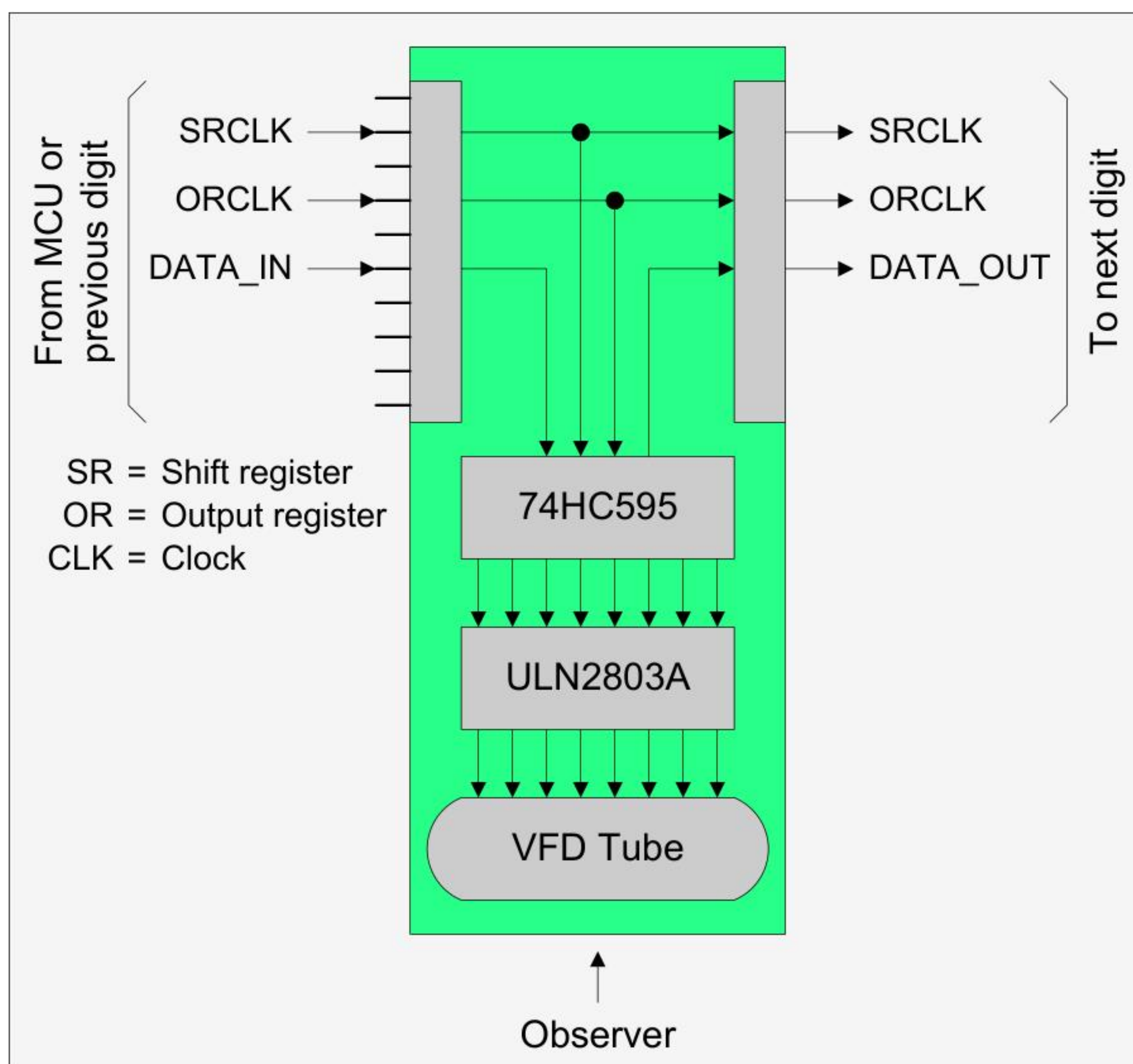


Fig.4: a high-level view of the functional flow.

your forehead, thinking: “VFD tubes are all well and good, but is life really worth living if tricolour LEDs aren’t involved?” Fear not—we have you covered. We’re going to pair a 5050 package (5 × 5mm) tricolour NeoPixel with each VFD.

Initially, we had planned to mount the NeoPixel on top of the board, directly beneath the tube. But experience has taught us to allow for the need to rework things if necessary. It would be somewhat embarrassing, frustrating, and time-consuming if a NeoPixel failed, and we had to go to all the effort of desoldering the 11-pin VFD just to replace it.

Our solution is even more cunning. In Fig.3, you’ll see a small rectangular cutout directly beneath the tube. This allows us to place the board face down, drop the NeoPixel (also face down) into the opening and spot-solder its terminals to pads on the underside of the board. This means we can replace the NeoPixel in the future, should that be necessary, without removing the tube.

A high-level view

Focusing on the VFD portion, a high-level view of the functional flow is shown in Fig.4. This is **not** a proper circuit diagram; it’s simply an abstract representation to give us a first-pass feel for how the main elements of the shift register chain and the data flow operate.

As we know from previous experiments, ‘595 shift registers actually contain two 8-bit registers: the shift register (SR) itself and the output register

(OR), which presents values to the outside world.

The **SRCLK** (shift register clock) signal loads the 0 or 1 value on **DATA_IN** into the shift register. Whatever 8-bit value is currently in the register shifts one bit to the right (where right is relative to the illustration in Fig.4). If multiple digits (boards) are daisy-chained together, the **DATA_OUT** from this board feeds into the **DATA_IN** of the next board in the chain.

Once all the bits in the shift register chain have been loaded, the **ORCLK**

(output register clock, sometimes called latch clock) signal is used to copy the contents of shift registers into the output registers, at which point they become visible to the outside world.

I use the term ‘MCU’ (microcontroller unit) in Fig. 4 rather than Arduino because our daisy-chain of boards can be driven by pretty much any microcontroller. We will be using an Arduino here, but Joe has opted to use a PIC microcontroller for his implementation.

I need some space!

If we were to simply connect 14 of these boards together to display my YYYYMMDDhhmmss values in one big ‘chunk’ (not wishing to be too technical), the result would be less than user-friendly. We need to space things out a little to look more like YYYY_MM_DD_hh_mm_ss.

Rather than using ribbon cables or similar connectors to connect groups of digits, we decided to create a special spacer board, illustrated in Fig.5.

At first glance, this board may appear to have nothing more than the two plug-and-socket connectors (boring). However, if you look closer, you’ll see a small rectangular cutout at the bottom of the board (exciting). This, of course, means that we will also be flaunting a tricolour NeoPixel on each of our spacer boards.

In turn, these pixels will allow us to implement separator characters between groups of digits, such as colons between the hh:mm:ss time fields. Thus, we don’t actually need a double space

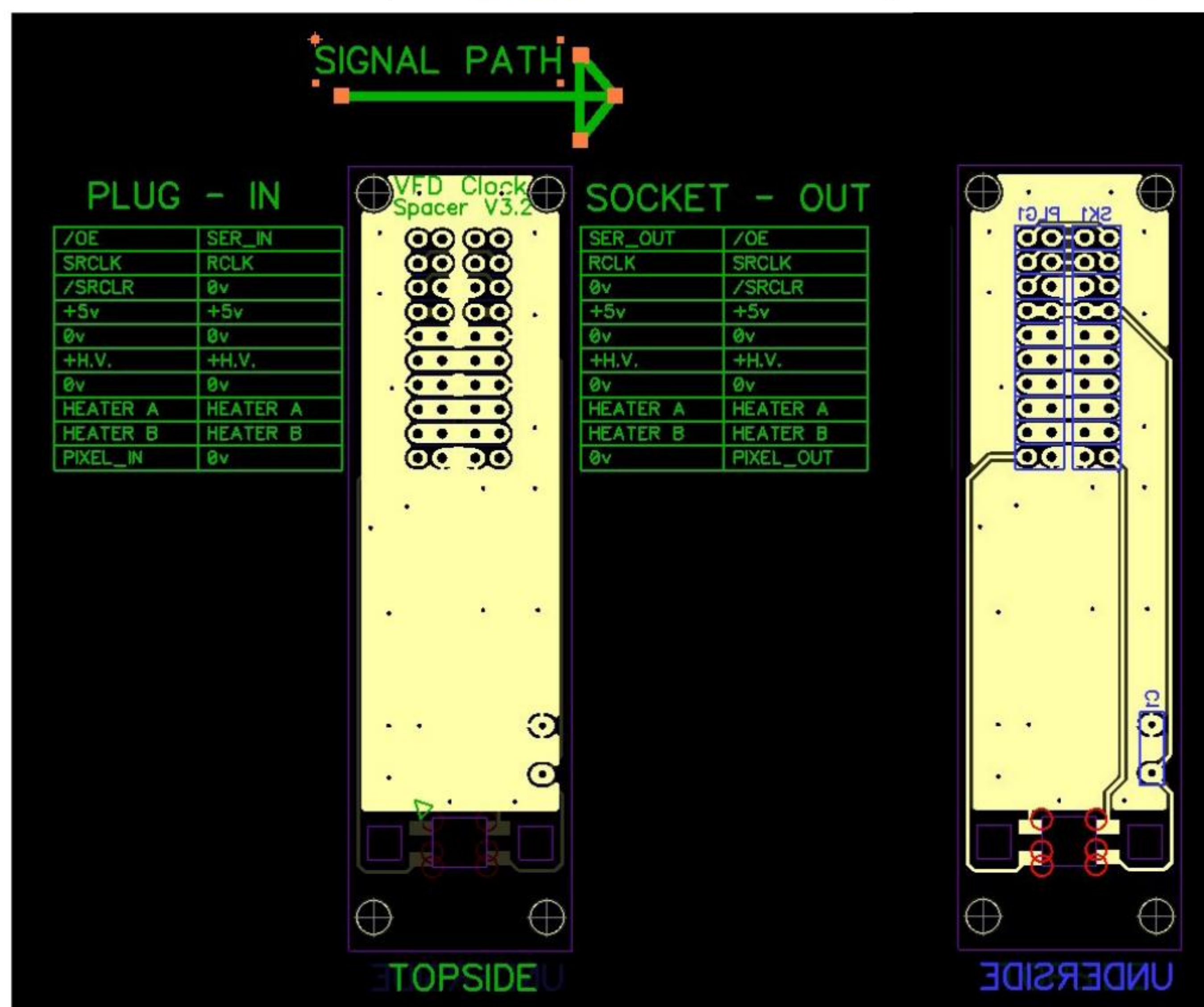


Fig.5: the layout of the spacer PCB; top side (left) and bottom side (right).



Photo 5: a first prototype of the Chronological Contraption by Joe Farr.

between the YYYY_MM_DD and the hh_mm_ss fields (although we could, if we wished, using two spacer boards) because we can simply leave this particular pixel off.

Joe already has an early prototype of the Cunning Chronometer up and running (I only just received my PCBs back from the fabrication facility). As seen in Photo 5, Joe's VFDs are fluorescent green, while the NeoPixels beneath his tubes are currently set to glow magenta.

Joe is using laser-cut, laser-etched acetate strips as separators, with the NeoPixels beneath them set to glow blue. Sad to relate, this didn't turn out quite as well as we'd envisaged, but we have more tricks up our sleeves (did you ever doubt us?) and we'll be experimenting with some tastier ways to do this in a future column.

Gabbing about Gerbers

The main way we convey information to a PCB manufacturing facility is via a file format known as Gerber, created by an extraordinary individual named H. Joseph Gerber.

There's a biographical book titled *The Inventor's Dilemma: The Remarkable Life of H. Joseph Gerber* ([amazon.com/dp/0300123507](https://www.amazon.com/dp/0300123507)), written by his son, David J. Gerber. As far as I'm concerned, this gripping tale falls squarely into the 'must-read' category.

Joseph became one of America's great inventors. Along the way, he revolutionised a variety of antiquated industries and created entirely new ones, including the digital plotters that transformed the way PCBs were manufactured.

These days, when we want to create something new, we can draw on a vast pool of existing infrastructure. Take the Arduino system, for example. If you want to use it to drive motors, there's a shield for that. If you want to connect your Uno to the internet via Wi-Fi, there's a shield for that, too. More importantly, someone else has already invented Wi-Fi... and the internet... for you.

This simply wasn't the case in Joseph's day; there was little in the way of infrastructure for him to rely on. As part of developing his digital plotters, he also had to define the file format for describing the plot data. Thus, the Gerber format was born.

Back to our boards

We've been through several iterations of these boards; we started out using IV-6 tubes, but they turned out to be too wimpy for our purposes. I still have a box of the little rascals sitting on the desk next to me in my office, making me feel guilty.

Joe has kindly made the Gerbers for the latest-and-greatest IV-11-based boards available to the rest of us (in a file named **CB-jun26-CC-gerbers.zip**). In addition to other files, this contains two more ZIP files: **Clock_IV11_V3.2_gerber.zip** for the main board and **Clock_Spacer_V3.2_gerber.zip** for the spacer board. You can simply upload these ZIP files to the PCB fabricator of your choice.

We'll also make these boards available from our website shop located at www.electronpublishing.com if you want to take the easy route.

I currently hang my hat in the USA, living amid a tumult of tariff turbulence that can transition day by day. While I have an honest desire to "buy locally", all the quotes I got from domestic PCB manufacturers made my eyes water. When I was informed my small order would be well over \$300 (£225), I soon found myself on the PCBWay website (www.pcbway.com).

This is one of the best-known low-cost PCB fabrication and assembly services used by hobbyists, startups

and even some professional engineers. They also support this publication, for which we are grateful!

This was the first time I've worked with these folks, and I have to say that I am very impressed. I uploaded the ZIP file containing the Gerbers for the main board and selected 20 boards (14 for starters and six spares 'just in case'). The cost of these boards was only \$21.81 (£16.50), which works out to just a tad over a dollar (~£0.83) per board. Now, that's more like it!

We should probably note that five boards is the minimum order quantity on PCBWay, but even if you need only one, think about this: one is for testing, one is for 'oops', one is for modifications, with two spares (or for friends).

Next, I uploaded the ZIP file containing the Gerbers for the spacer board and selected 10 boards (five for starters and five for spares). I know these are smaller and simpler than the main board, but I was still surprised to see a price of \$5 (~£3.80) for all 10 boards, which even my poor mathematical skills tell me works out to only \$0.50 (~£0.38) a board.

The total came to \$26.81 for all the boards (about £20), which brought a cheesy smile to my chubby cheeks. Even after adding \$27.27 for shipping





and \$22.47 for assorted extras (bank fees, duties, taxes and tariffs), I was more than happy with the grand total—still only around \$2.50 (£1.90) per board.

Just a couple of weeks later, I had my Cuning Chronometer boards in my eager hands, and they look awesome (happiest of happy dances). We will move on to consider the other components next month, but first...

Ask not for whom...

The phrase “for whom the bell tolls” originates from a 1624 meditation by John Donne, who wrote, “never send to know for whom the bell tolls; it tolls for thee”. Ernest Hemingway later borrowed the line as the title of his 1940 novel *For Whom the Bell Tolls*, helping to popularise the form of the saying that most people quote today.

I’m mentioning this here because I’ve just remembered one facet of the Cuning

Chronometer I wanted to share with you. When the count reaches zero (well, 0000 00 00 00 00 00)—at 11:45am British Summer Time (BST), thereby marking the 100th anniversary of my arrival on this planet—the Cuning Chronometer will burst into a veritable cornucopia of exuberant sound effects.

I’m thinking a carillon pealing forth in exultation would make a rather nice touch.

I mentioned my belief that my Cuning Chronometer is destined to become a treasured family keepsake. Thus, even if I’m not there to witness the completion of the countdown myself (I may be off canoeing up the Amazon or something of that ilk), my son Joseph and his family can gather round to tell each other tall tales of the magnificence that is Max.

Of course, this may start to wear a little on the nerves after the first couple

of hours, at which point I expect someone to (reluctantly) unplug the unit from the wall. Unfortunately (for them), they will be blissfully unaware of the ultra-top-secret backup power supply embedded within the device.

I think it would be a nice touch if the sound of the bells tolling were to return around 3:00am. the following morning, at which point we will all know “for whom the bell tolls”.

Next time

We will commence our next column by adding a 9DOF IMU to our retro game console. Next, we will delve deeper into VFD technology, consider the components required to implement the Cuning Chronometer, and start assembling its boards.

As always, if you have any thoughts you’d care to share on anything you’ve read here, please feel free to email me at max@clivemaxfield.com **PE**

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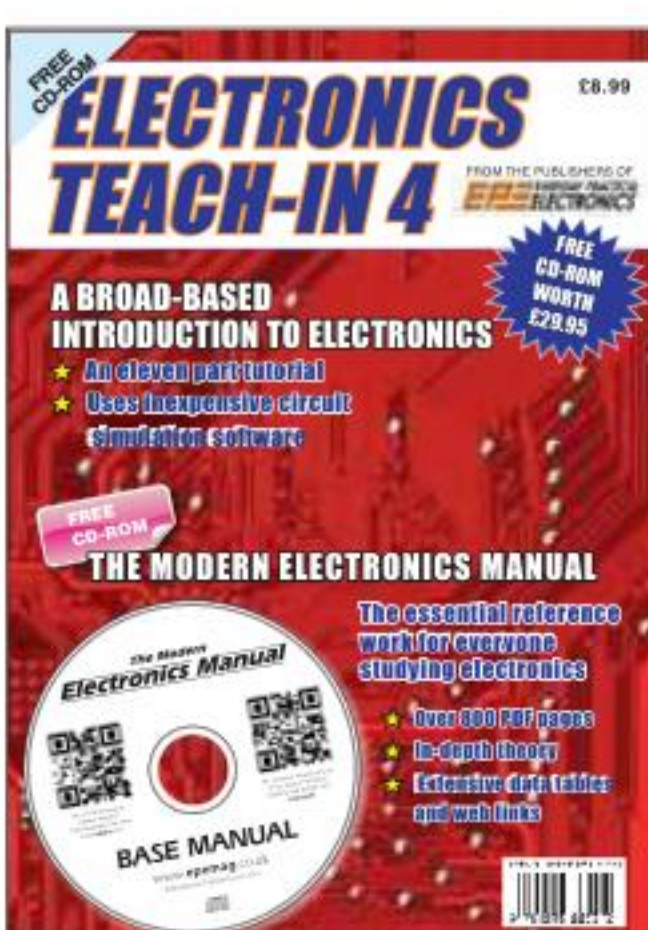
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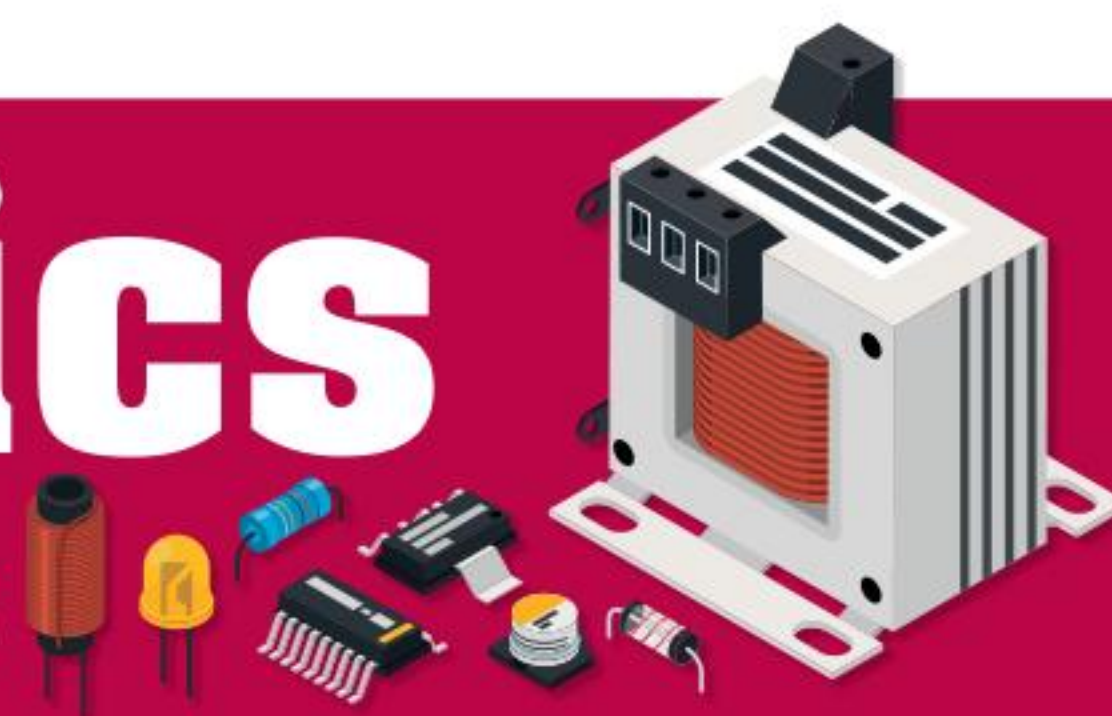
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By Andrew Levido

Power Electronics

Part 3: Isolated DC-DC Converters



Isolated DC-DC converters offer both improved safety and flexibility compared to the non-isolated kind. To design an isolated converter, first we need to understand how transformers work, plus their potential roles in DC/DC and AC/DC converters. This article will start with transformer/inductor fundamentals, then move on to using them.

The DC-DC converters we have looked at so far have been non-isolated types.

That means there is a direct electrical connection between the input and output. In many cases, we want the output to be isolated from the input; for safety reasons, if the input is connected to the mains, or because we need the output to be referenced to a different potential than the input.

Isolation is usually achieved using a transformer. Adding a transformer to a switching converter can provide a host of other benefits – for example, it can reduce the range of duty cycle (and consequent component stress) required to achieve high step-up or step-down ratios. It can also allow us to get multiple outputs from a single converter.

As we have seen in this series so far, we cannot get very far in the field of power electronics before coming up against magnetic components such as inductors and transformers. Like everything else, it is the second-order non-ideal behaviour of these components that will catch us out if we are not careful.

Transformers, in particular, are highly specific to the application, so sooner

or later we will have to roll our own. This means that a solid understanding of magnetics theory is necessary before we get to isolated DC-DC converters, so let's dive in.

Back to basics

One of the reasons magnetics can be confusing is the number of terms that sound similar but have different meanings. This comes about because magnetics was one of the earliest areas of electrical engineering to be studied by the likes of Gauss, Ampère and Faraday, who each invented their own terms, which we live with to this day.

With terms like magnetic field intensity, magnetic flux, magnetic flux density, magnetic flux linkage, magnetic permeability and magnetic permeance, it is no wonder many of us get confused. Don't even get me started on Maxwell's equations! (We will have an article on those equations in a future issue.)

I promise that it is not as hard to wrap your head around as you might think. As usual, I will not cover this topic in a rigorous academic fashion, but from the perspective of a working engineer. We will employ just a little basic algebra.

The left side of Fig.1 shows a winding of several turns of wire around a magnetically permeable core (we'll consider what that means soon). The N-turn winding carries a current i , which induces some kind of magnetic flux (or field in some textbooks) denoted by the Greek capital phi (Φ), shown in red. The core itself has a cross sectional area A and an average length l .

By convention, the direction of the flux follows the 'right-hand rule' illustrated at the bottom left of the figure. If you curl the fingers of your right hand in the direction of current flow, the resulting flux points in the direction of your thumb.

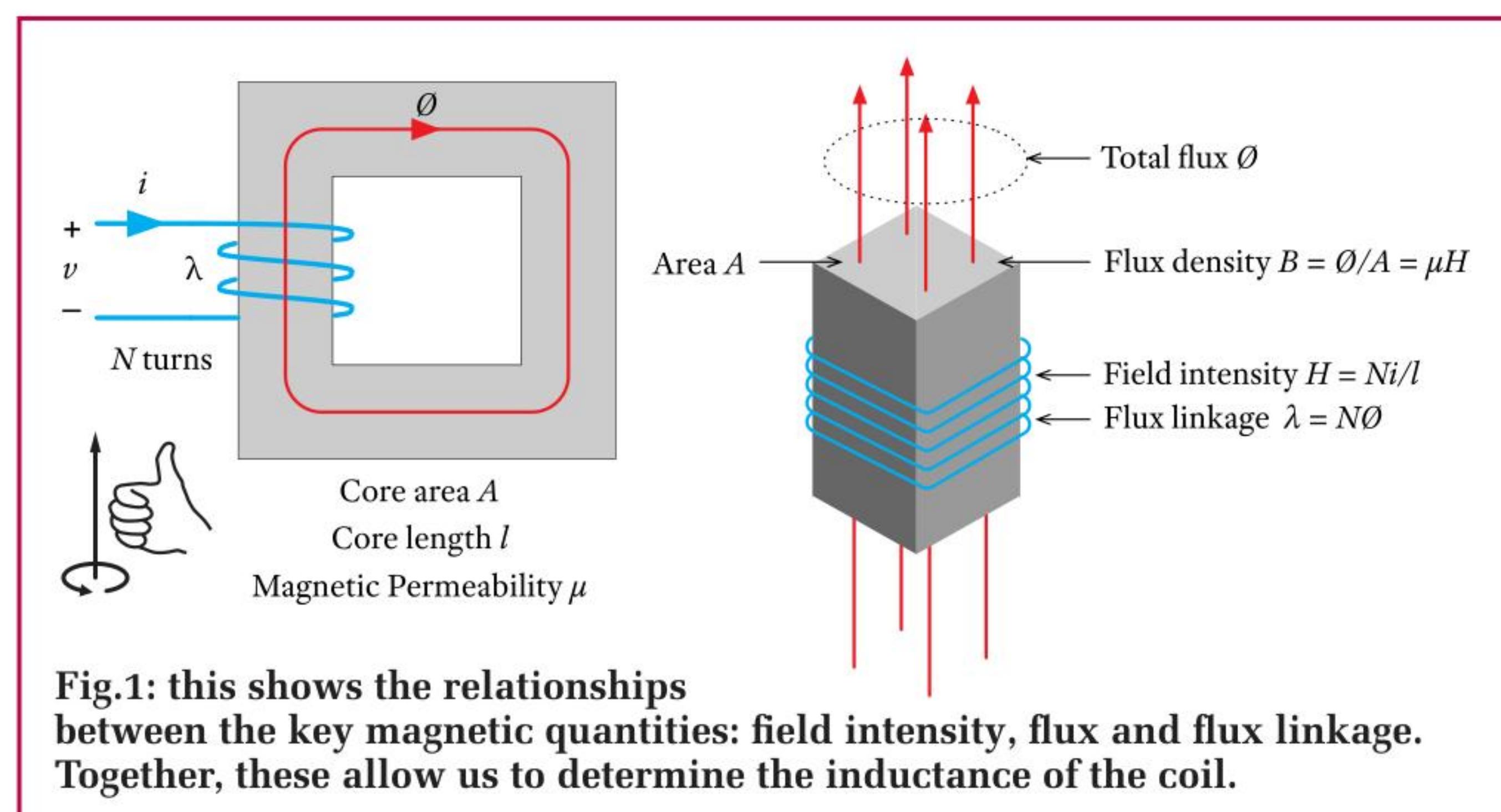
Gauss' law for magnetism tells us that the net flux entering and leaving any closed region is zero, which means that flux lines have no beginning or end – they must always be closed loops. For us, this just means that all of the flux leaving the coil at the top enters it again at the bottom. For now, we will assume that all of the flux is confined to the core.

You can think of this arrangement in two ways at the same time; as loops of current enclosing a flux, or as loops of flux enclosing a current.

Flux is driven around the core by a 'driving force' known as the magnetomotive force (mmf) that describes the amount of current linking with the flux. This mmf is N times the current i , since the current passes through the flux loop once for each turn of the winding.

We can also describe this magnetising force as a magnetic field intensity, which we denote with the letter H . This is defined by Ampère's law to be equal to $Ni \div l$, or the mmf per unit length of core.

The right-hand side of the figure shows a section of the core with the flux lines passing through it. As we



Part 3: Isolated DC-DC Converters

have seen, the magnetic field intensity H drives a total flux Φ through the core, equal to $\mu H A$. This results in a flux density, B , which is simply the amount of flux divided by the cross-sectional area of the core.

B is also related to H by the magnetic permeability, or μ , of the core according to the relationship $B = \mu H$. Permeability is a measure of how 'easily' a given field intensity can create a flux density in the core. Materials with a higher permeability will develop a higher flux density for a given field intensity.

The permeability of a material is usually expressed as the permeability of free space, μ_0 (which is equal to $4\pi \times 10^{-7} \text{H/m}$) multiplied by a unitless relative permeability, μ_r .

The permeability of free space is the basic measure of how much flux a given current will produce in a vacuum or in air. The relative permeability is a measure of how many times more permeable a material is than this baseline. Thus, you often see permeability expressed as $\mu_0 \times \mu_r$.

Relative permeabilities for magnetic materials range from a few hundred for mild steel, to 5000-10,000 for transformer steel, up to 40,000 or more for some ferrites.

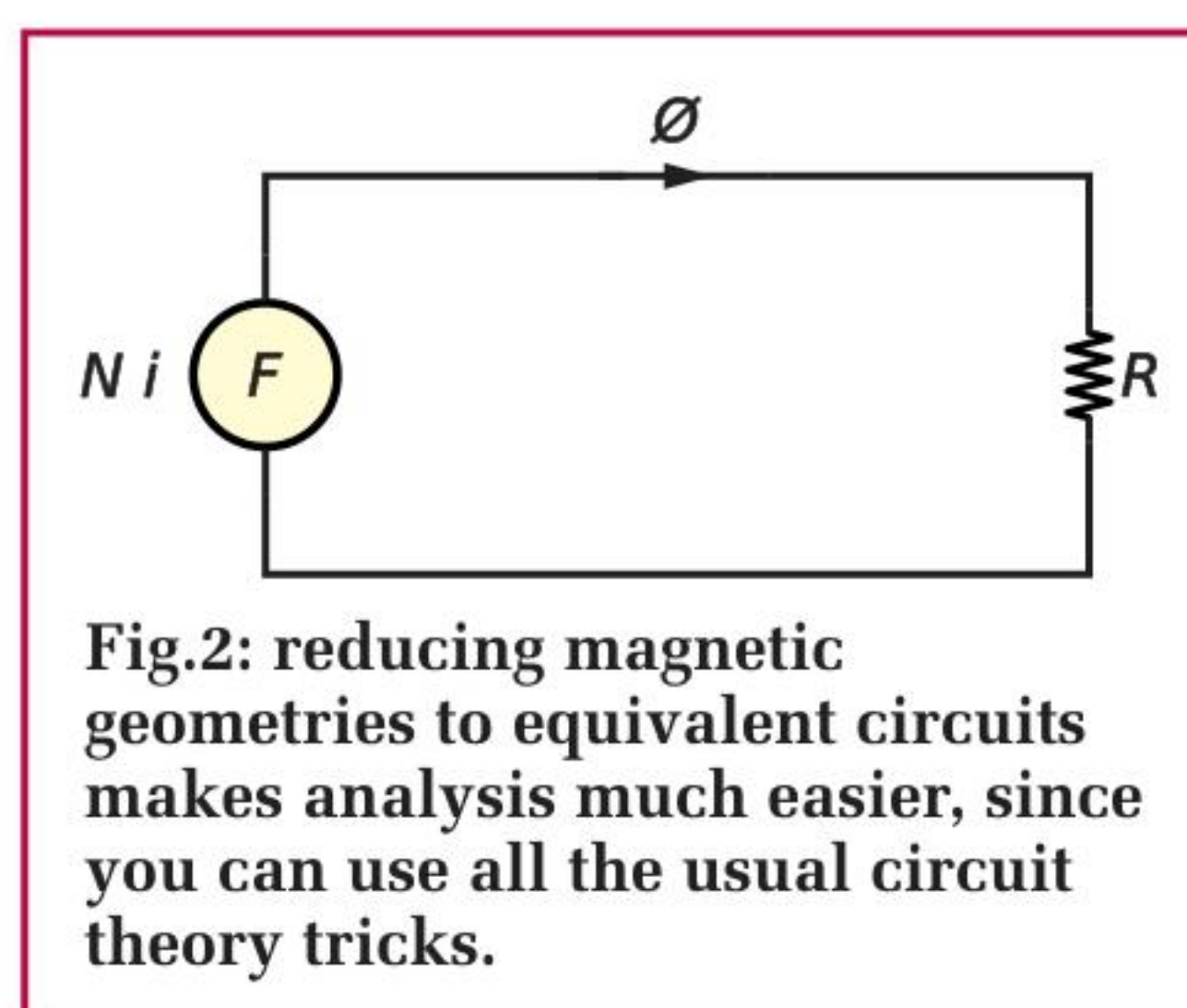
In addition to the flux and the flux density, we need to introduce the concept of flux linkage, which describes how a flux links with a winding. From Fig.1, you should be able to see that the red flux lines pass through each turn of the winding, giving it a flux linkage $\lambda = N\Phi$ (λ is the Greek letter lambda).

So, to sum up, the current flowing in a winding produces a magnetic field intensity, which drives a flux around the core. This produces a flux density in the core that is related to the field intensity by the permeability, and to the total flux by the cross-sectional area of the core. The resulting flux passes through the winding, resulting in a flux linkage, λ .

Inductors

You might have noticed that this is a bit circular – the current in the winding produces a flux that links with the winding. By introducing the third member of the magnetic holy trinity, Mr Faraday, we can use this to understand how inductors work.

Faraday's law states that a voltage is induced in a winding by a changing magnetic field. In our example, the



magnetic flux will change proportionally to changes in the current, so it follows that a voltage is produced across the terminals of our winding as the current through the winding changes.

The changing coil current effectively induces a voltage in itself. This is known formally as self-inductance, but we usually just refer to it by the shorter name "inductance". Inductance kind of wraps up all of the N , H , B , Φ , and μ malarky into a relationship between the current and the flux linkage. In fact inductance, is defined as the flux linkage per unit of current, $L = \lambda \div i$.

For the mathematically inclined, you can see how this works by working out an expression for the inductance of the arrangement in Fig.1. Combining the equations for field intensity ($H = Ni \div l$) and flux density ($B = \mu_0 \mu_r H$) we get $B = \mu_0 \mu_r Ni \div l$. Multiplying by the area gives us the total flux $\Phi = \mu_0 \mu_r Ni A \div l$.

We can then use the formula for flux linkage ($\lambda = N\Phi$) to get $\lambda = \mu_0 \mu_r N^2 i A \div l$, and finally use the formula for in-

ductance to arrive at $L = N^2 \times \mu_0 \mu_r A \div l$. The inductance is therefore the product of the square of the number of turns multiplied by a term related to permeability of the core and its dimensions.

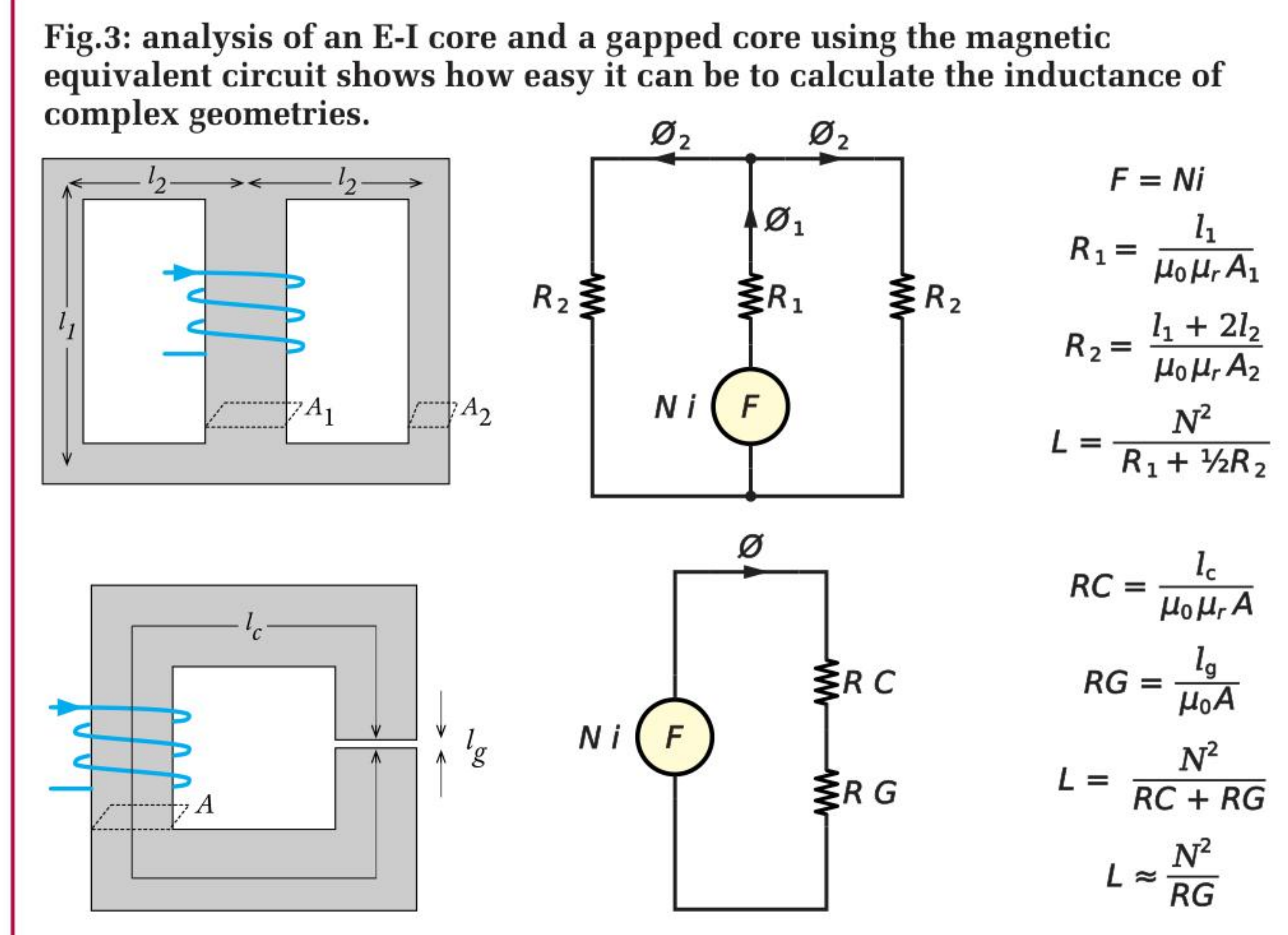
This latter term is referred to as the permeance of the core (not to be confused with the permeability). If you know the permeance of a core, you can easily calculate the number of turns required to obtain a given inductance. Manufacturers of cores usually provide the permeance in their data sheets as an " A_L " value, in units of nanohenries per turn squared or similar.

Electric circuit model

The inverse of permeance is reluctance, which is an extremely useful quantity we can use to build a model of magnetic systems that is analogous to electric circuits. In these circuits, reluctance (denoted R) is equivalent to resistance and the mmf (F), equal to Ni , is equivalent to voltage. The resulting flux (Φ) is analogous to current.

Fig.2 shows the electrical circuit equivalent of the magnetic circuit in Fig.1. The circuit obeys the magnetic version of Ohm's law, so $F = \Phi R$.

We can use all of our usual circuit analysis techniques, so this is really helpful to calculate inductances and the like when faced with more complex core geometries such as that in Fig.3. At the top, we have a classic E-I core with a winding on the centre leg that is wider than the outer legs.



The magnetic equivalent circuit is shown to the right. It is easy to calculate the reluctance of the centre leg, R_1 , and that of the two outer legs, R_2 , based on the dimensions of the core and its permeability. You can then use what you know about resistors in series and parallel to calculate an equivalent reluctance and therefore the inductance as shown in the figure.

Fig.3 also shows another common configuration, where a very narrow air gap is included in the magnetic circuit. This is done to increase the stability of the inductance and the amount of energy that can be stored in the core.

The reluctance of the air gap is much higher than that of the core, because the relative permeability of air is one, compared to many thousands for the core. This means the inductance is dictated by the air gap, and is largely independent of the core material.

This can be a good thing for the stability of the inductance, since the relative permeability of most core materials changes with temperature and flux density, as we will see below.

I won't cover the maths, but for the same reason, for a given flux density, the amount of energy that can be stored per unit volume is much higher in the air gap than in the core.

In fact, it is common to assume that all of the energy is stored in the gap, and it's not unusual to start the design process for inductors by selecting a core with sufficient gap volume (the area of the core times the gap length) to store the required energy each switching cycle.

Leakage

As usual, I have made a few simplifications in the above discussion, and some other factors come into play when we get into the nitty-gritty of magnetics design. One of these is leakage flux.

In the above examples, we assumed that all of the flux was constrained to the core.

Fig.4 shows that this may not be the case – some flux may leak away from the core and pass through the air where there is a lower reluctance path; however, it will always return, due to Gauss' law.

The effect of this on the magnetic circuit is shown in the right-hand side of the figure. The leakage paths form a leakage (generally high) reluctance, which appears in parallel with the core reluctance. You can see from the formula that this leakage will result in a slightly increased inductance over the ideal case.

It will be useful later to think of the leakage producing an extra 'leakage inductance' in series with the main core inductance, as shown in the lower equation.

Saturation, hysteresis and residual flux

We have also assumed up until now that the relationship between field intensity and flux density is linear, described by a simple proportional relationship of permeability. The reality (as always) is a little more complex. A typical magnetic material has a B-H characteristic like that in Fig.5, although it is shown a bit exaggerated for clarity.

Remember that H is the magnetising 'drive', proportional to the winding current, and B is the resulting flux density. There are three important things to note. First, the path that B follows when H is increasing is not the same as it does when H is decreasing; there is hysteresis.

Second, neither path passes through the origin. When H is zero, there will be some residual flux density B_{res} (either positive or negative) present in the material when it is not excited. Thirdly, the slope of the characteristic (the per-

meability) saturates at some flux density, B_{sat} – it deviates from the ideal value of μ shown in red as the field intensity increases.

We normally want to avoid saturation (although there are some notable exceptions where this characteristic is used to advantage), so we take some care to ensure the maximum flux density stays well below the saturation level.

Losses

We have also so far assumed inductors are lossless, but of course, we know this cannot be the case. Losses in magnetic components fall into two categories: copper losses (sometimes called winding losses) and core losses.

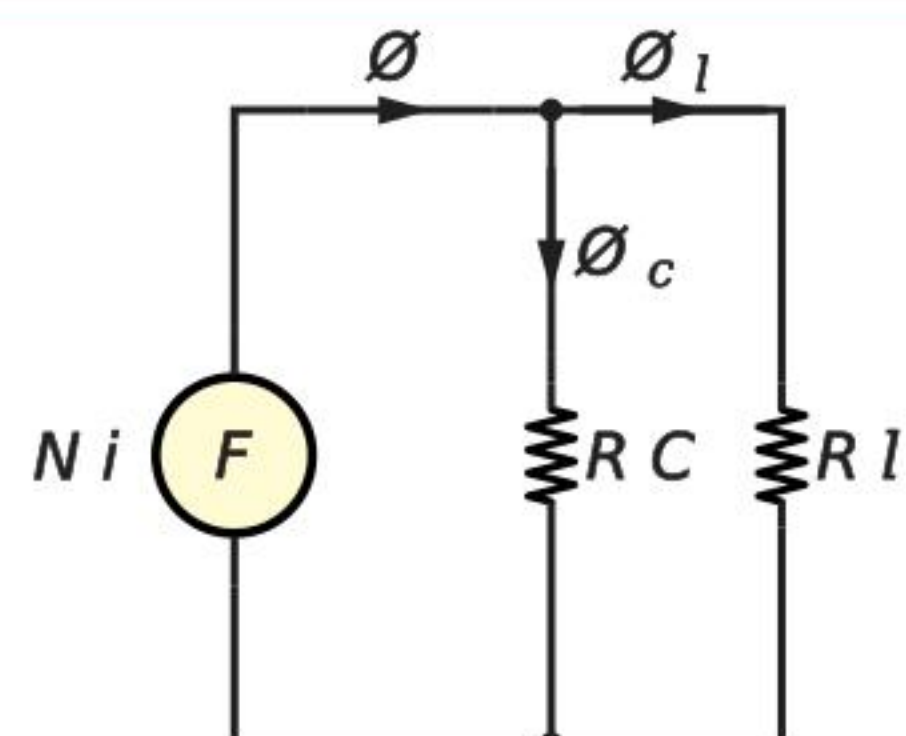
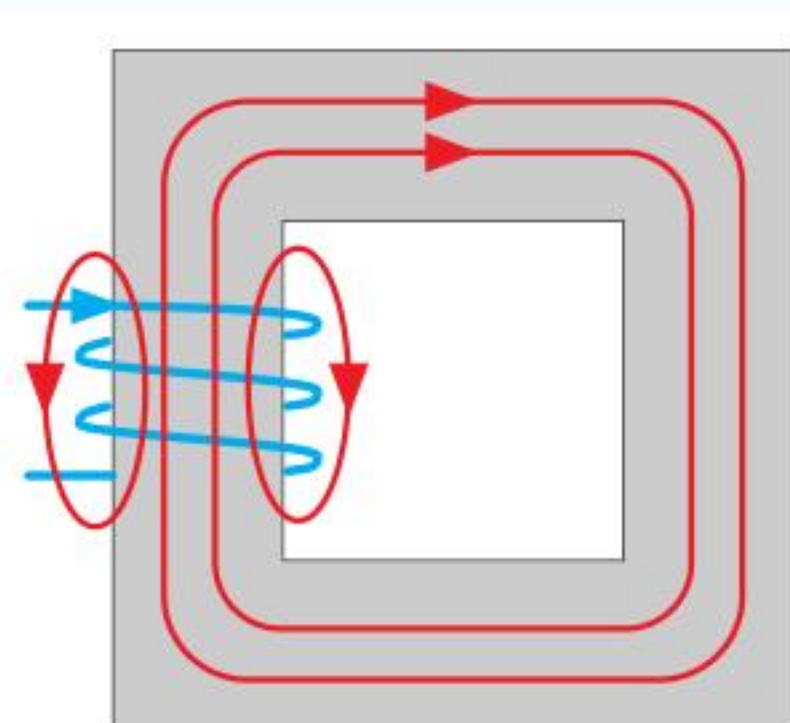
Copper losses include the familiar ohmic loss determined by the resistivity of the winding material, its cross sectional area and its length.

Be aware that the resistivity of copper, the most common winding material, is temperature-dependent and increases by approximately 40% for every 100°C temperature rise. Make sure to calculate losses using the resistivity at the highest operating temperature you will see in the winding.

Resistive losses are exacerbated by a phenomenon known as 'skin effect'. As the frequency of a current passing through a conductor increases, eddy currents create magnetic fields inside the conductor that force the current outward, so it flows only in the outside 'skin' of the conductor. The higher the frequency, the more the current is forced to the outside of the conductor.

The 'skin depth' is a measure of how much of the conductor is effectively useful.

For copper, the skin depth is about 9.2mm at 50Hz, which explains why high-current AC busbars tend to be



$$L = N^2 \left(\frac{1}{R_C} + \frac{1}{R_L} \right)$$

$$L = L_C + L_L$$

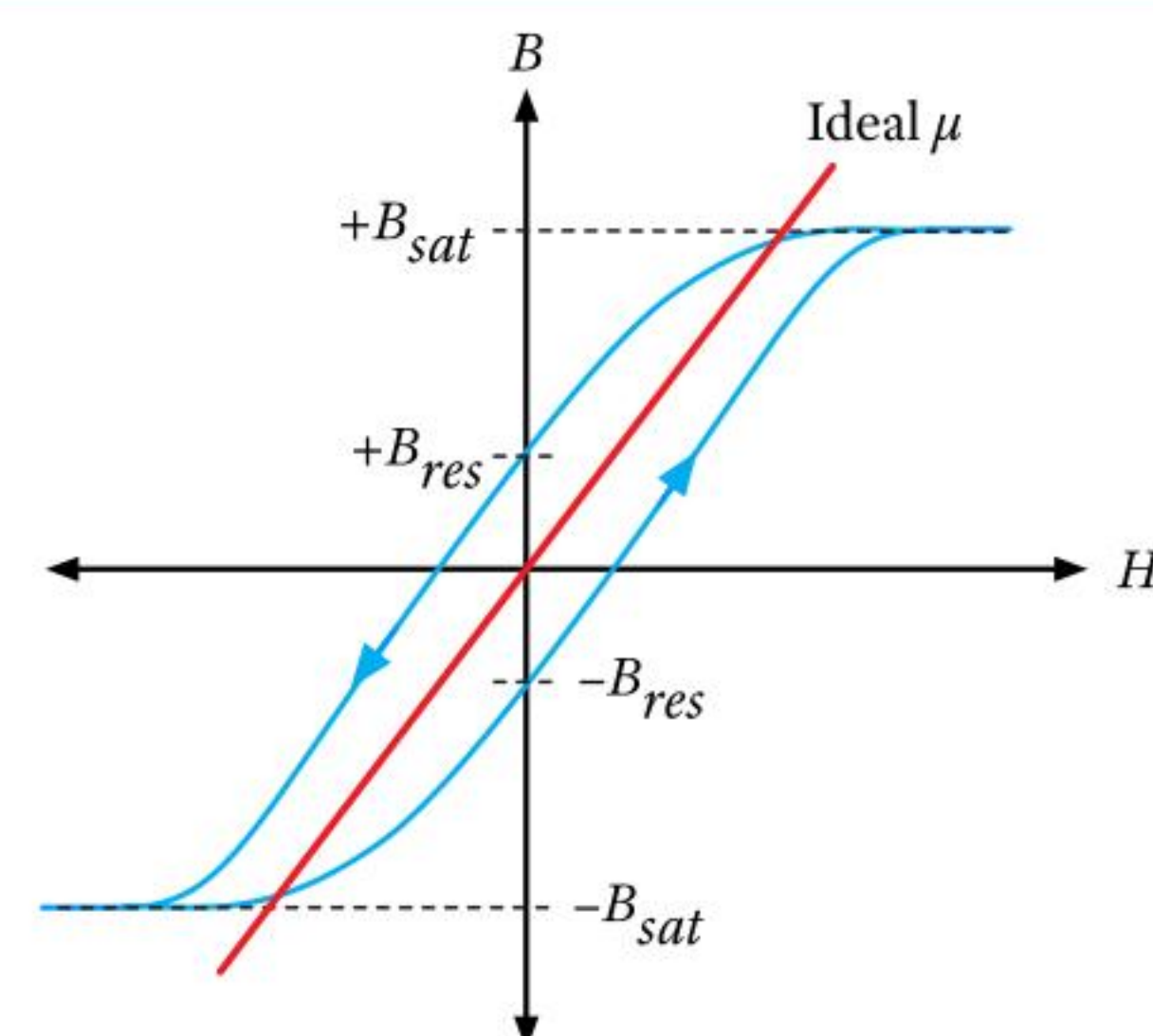


Fig.4: flux leakage in an inductor produces an extra leakage reluctance in parallel with the core's reluctance, and increases the total inductance slightly.

Fig.5: magnetic materials have a less-than-ideal B-H characteristic that includes hysteresis, saturation and residual flux density.

Part 3: Isolated DC-DC Converters

broad but are rarely more than 10mm thick. At 100kHz, the skin depth is about 0.2mm, and at 1MHz, it is just 65µm. There is no point in using a cylindrical conductor with a diameter greater than twice the skin depth, since there will be no corresponding reduction in resistance.

At 100kHz, for example, any conductor with a diameter larger than 0.4mm will be a waste of copper.

For this reason, the magnetics in high-power switching converters use multiple thin conductors in parallel, or are wound with copper foil (or even multiple parallel layers of copper foil). You can also use Litz wire, which is an intricate arrangement of fine insulated wires twisted together into bundles, which are themselves twisted together. It is lovely stuff, but expensive.

Core losses also come from two sources: eddy currents and hysteresis. Eddy current losses are resistive losses caused by currents circulating within the core material. In conductive core materials like steel, these losses are mitigated by making the core from thin laminations that are insulated from each other by an oxide layer.

You will have seen these laminations in the cores of E-I type mains transformers. Toroidal mains transformer cores are wound from a long thin strip of steel (like a roll of sticky tape) to achieve the same end. At higher frequencies, we tend to use cores made of materials that have poor electrical conductivity, such as ferrite or sintered metal oxides, to avoid eddy current losses.

Hysteresis loss is caused by the shape of the B-H curve. When the material is magnetised in one direction, it takes some magnetic force in the other direction to overcome the residual flux and bring the flux density back to zero. This takes energy, which becomes heat in the core.

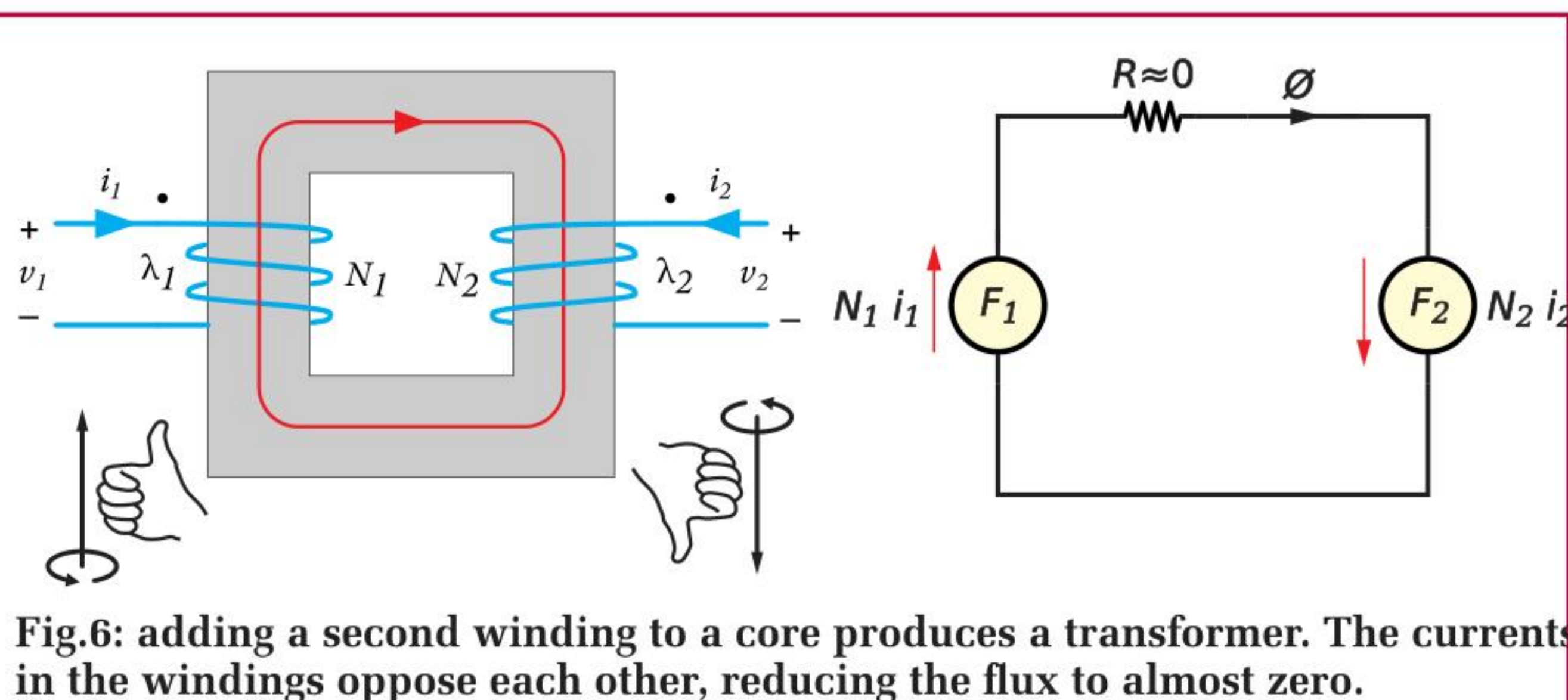
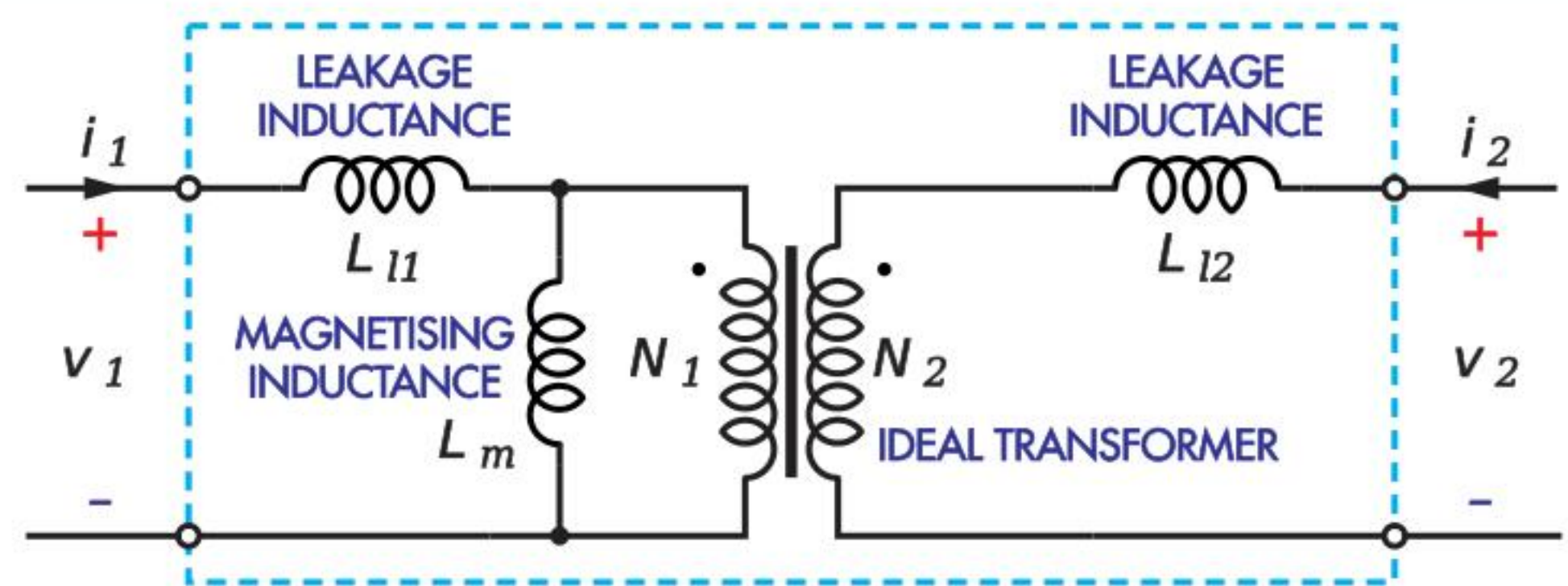


Fig.6: adding a second winding to a core produces a transformer. The currents in the windings oppose each other, reducing the flux to almost zero.

Fig.7: a realistic transformer has leakage inductances due to imperfect coupling of the flux, and a magnetising inductance due to the finite permeability of the core.



The amount of loss is proportional to the area within the hysteresis loop, so choosing a material with a narrower hysteresis curve will help minimise these losses, as will limiting the maximum flux density excursions.

Manufacturers usually provide a measure of core losses for their materials in kW per cubic metre for a (usually pretty small) range of frequencies and flux densities in the data sheets. You have to multiply these by the core volume to get an estimate of core loss in your application.

Transformers

Adding a second winding to our inductor, as shown in Fig.6, produces a transformer. If the flux is perfectly linked by both windings, as shown here, the transformer is said to be perfectly coupled. While we are at it, let us also assume that the core is so permeable that the reluctance is zero. This is, after all, an ideal transformer.

Since the flux linked by each turn on both windings is identical, so is the voltage produced across each turn. Each winding voltage is therefore proportional to the number of turns in that winding. The ratio of voltages $v_1:v_2$ is equal to the turns ratio, $N_1:N_2$.

Due to the sense of the windings in the diagram and the right-hand rule, the total mmf is the sum of the Ni values of each winding. If current flows into the dotted ends of either winding, it produces a clockwise flux as shown. I have indicated the direction of flux for

a current into the dotted terminal by a red arrow on the electrical equivalent circuit on the right.

The magnetic circuit shows that even though the two mmfs are pushing flux around the core in the same direction, when one is excited, the other will see an mmf of the opposite polarity. The mmf seen at F_2 due to a positive F_1 will be negative and vice versa. This means that a current entering the dotted terminal of one winding will force a current out of the dotted terminal on the other winding.

If the reluctance is zero, the mmfs will be equal in magnitude as well as opposite in sign. In other words, the ratio of transformer currents $i_1:i_2$ is equal to $-N_2:N_1$. Don't get too worried about the negative sign here; it is just there because convention says positive current flows into the dotted terminal.

In an ideal transformer, then, perfect flux linkage means the voltages are related by the turns ratio ($v_1:v_2 = N_1:N_2$), and zero reluctance means the currents are related by the inverse of the turns ratio ($i_1:i_2 = N_2:N_1$). The net flux in the core must be zero since the input and output mmfs cancel out as they are identical but opposite in sign.

The impedance looking into one winding with the other open will be infinite (it will look like an open circuit), and when the other is short-circuited, it will be zero.

A transformer model

While ideal transformers are handy for circuit analysis, they are not realistic. We can summarise these non-idealities in the equivalent circuit of Fig.7. There is an ideal transformer in the centre of the diagram. The inductances in series with each side are leakage inductances caused by incomplete coupling of the flux, as we saw in Fig.4.

I have shown a leakage inductance on either side of the transformer, but it is also sometimes handy to have it all 'lumped' onto one side or the other.

For example, if we wanted to show L_{l2} on the same side as L_{l1} , we would shift it over but multiply its value by $(N_1 \div N_2)^2$. We could equally move L_{l1} to the same side as L_{l2} by multiplying its value by $(N_2 \div N_1)^2$.

Since any real transformer core has a finite reluctance, the opposing mmfs of each winding will not completely cancel out, and there will be some level of residual flux in the core. This is represented by the parallel inductance known as the magnetising inductance. This is responsible for the small current that will flow in a transformer's primary winding when its secondary is open circuit.

There are also copper losses and core losses in transformers, driven by the same mechanisms as discussed above for inductors. The copper losses can be represented by appropriate resistances in series with the leakage inductances, and the core losses by a resistor in parallel with the magnetising inductor. I have not bothered to show them here to keep things simple.

The forward converter

Knowing what we do about magnetics, we can begin to understand isolated DC-DC converters. In the upper-left corner of Fig.8 is a non-isolated buck converter that we are by now very familiar with. To its right is an isolated version of the same topology, known as a single-ended forward converter.

An ideal (for now) transformer with a turns ratio of $N:1$ has been inserted pretty much in the middle of the buck converter's switch $Q1$, which now consists of a Mosfet ($Q1$) on the primary side and diode ($D2$) on the secondary side.

When $Q1$ is on, current flows into the dotted primary terminal of the transformer. A current, scaled by a factor of N , emerges from the dotted secondary terminal and passes through $D2$, forming the second half of the switch feeding the filter inductor. When $Q1$ and $D2$ are off, the filter inductor current flows via diode $D1$, just as it does in the non-isolated converter.

The transfer function of the forward converter is the same as the buck converter, but scaled by the transformer turns ratio, N . I have drawn the forward converter with the Mosfet in the positive input line to match the buck converter, but in reality, it is typically moved to the 'ground' side of the transformer primary to make its gate drive simpler, as shown in the circuit at lower left in Fig.8.

In this circuit, I have also added the magnetising inductance to the transformer, and a clamp circuit consisting of diode $D3$ and zener diode $ZD1$. You can probably already see why these are necessary. We don't need to worry about the leakage inductances, as they are in series with the ideal transformer, so there is always a path for their current to flow.

They will affect voltage regulation a bit, but we won't worry about that now.

This is a 'single-ended' converter because power flows through the transformer only during the part of the cycle when $Q1$ is conducting. This means that when the Mosfet switches off, the magnetising current needs a path to flow or else the Mosfet's drain terminal voltage will spike and it will be toast.

The clamp circuit limits the Mosfet drain voltage to the sum of V_{IN} plus the zener voltage. The energy stored in the magnetising inductance is dissipated in the clamp every cycle.

The price we pay for the convenience of the transformer is the additional complexity and power dissipation of a clamping circuit, and the additional voltage stress on the switch. There are several other ways to implement the clamping circuit, two of which are shown at lower right in Fig.8.

The first, a resistor-capacitor-diode (RCD) clamp, relies on a capacitor to absorb the energy, which is then dissipated in the parallel resistor. This is probably the cheapest option and is often seen in low-cost designs.

A more efficient option is to add an extra winding to the transformer and a diode, as shown in the energy recovery clamp partial circuit. When the Mosfet switches off, the magnetising current causes the transformer terminal voltage to rise until the clamp diode conducts.

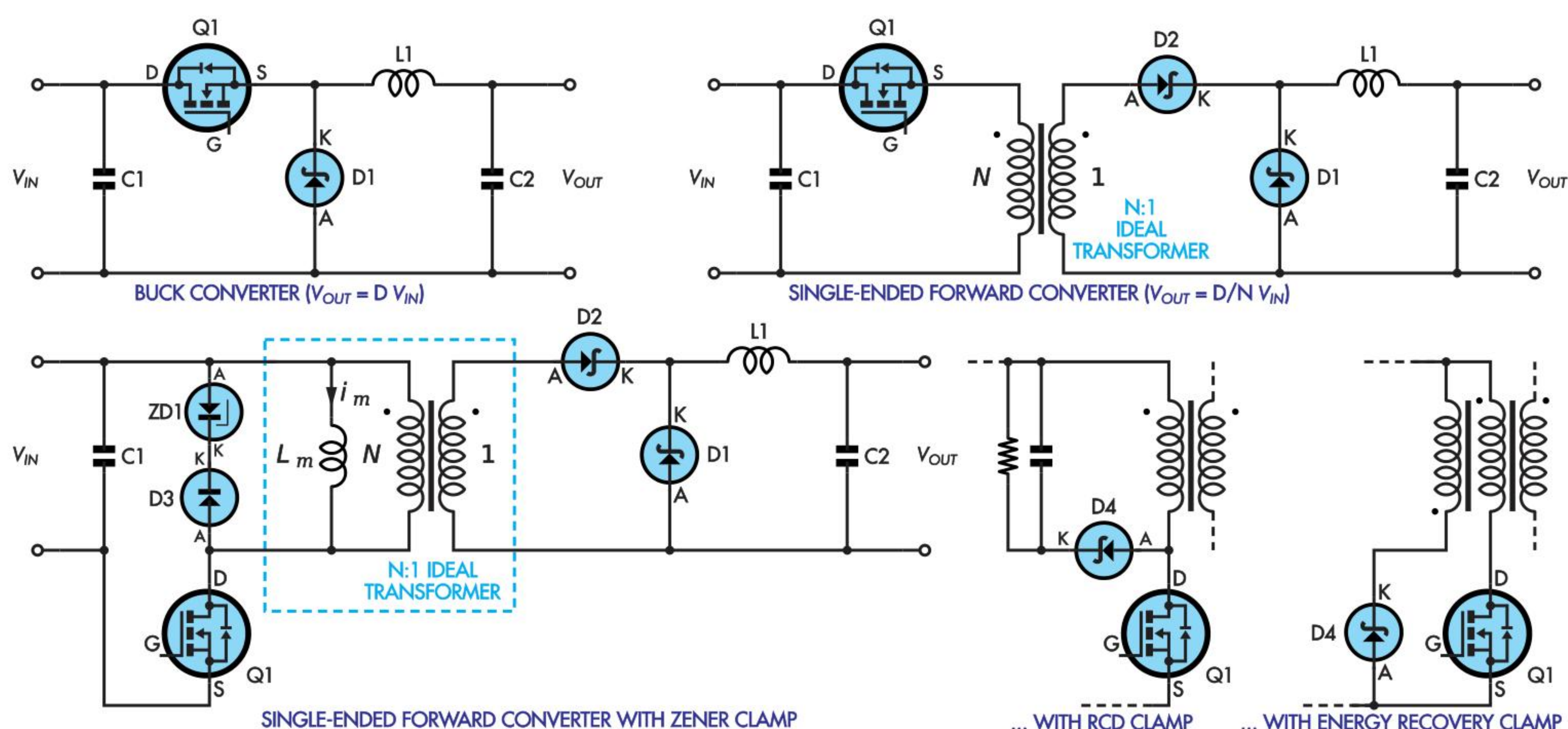


Fig.8: the single-ended forward converter is essentially a buck converter with $Q1$ replaced by a Mosfet, transformer and diode. The transformer's magnetising inductance requires the addition of a clamp circuit, as shown along the bottom of the figure.

Part 3: Isolated DC-DC Converters

If the clamp winding had the same number of turns as the primary, the Mosfet drain voltage would be clamped to twice V_{IN} . You can choose the number of turns on the clamp winding to limit the drain voltage even further if necessary. As the name implies, the advantage here is that the magnetising energy is returned to the supply.

Another similar variant of the single-ended forward converter, the isolated hybrid bridge converter, is shown in Fig.9. It is still a single-ended converter, because the Mosfets can only drive flux through the transformer in one direction, but it solves the magnetisation current problem by clamping both ends of the windings to the supply rails when the Mosfets are off.

The Mosfets are never subject to more voltage stress than the supply rails, but the gate drive for the upper Mosfet is more complex in this configuration.

Double-ended forward converters

An obvious(?) next step would be to replace the diodes in the hybrid bridge converter with Mosfets to produce a full-bridge converter. This has the huge advantage of driving flux in both directions in the transformer and allowing us to use a full-bridge rectifier on the secondary side.

The single-ended converter can only drive the magnetising flux around the transformer core in one direction (re-

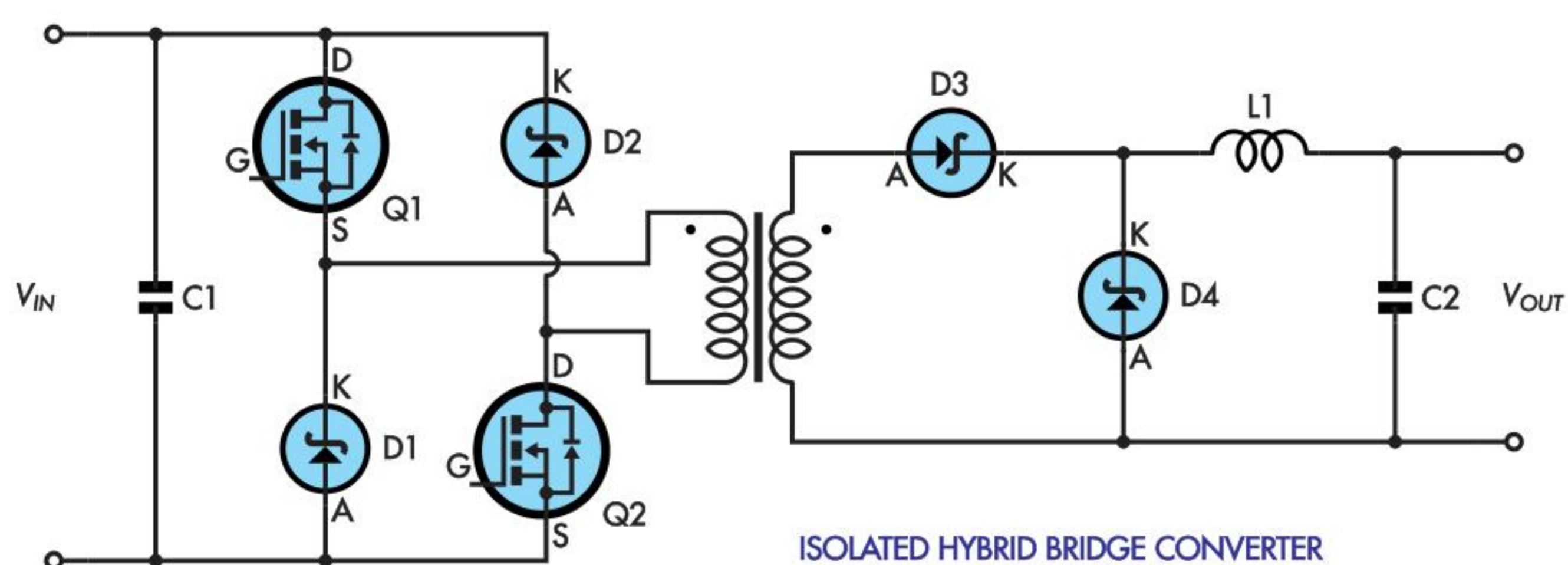


Fig.9: the isolated hybrid bridge converter solves the problem of magnetising inductance at the cost of circuit complexity.

member, the magnetising flux is what's left in the core after most of the flux is cancelled out). In a double-ended converter such as this one, the magnetising current can change sign. We can therefore utilise the full range of flux density in the core material, making for a more efficient and smaller transformer.

Being able to use a full-bridge rectifier means we can use a smaller filter inductor, since the frequency at the output of the rectifier is twice the switching frequency. This type of converter does require that we take care to limit the duty cycle of each phase to less than 50%, or we run the risk of switching on the upper and lower Mosfets at the same time, with catastrophic results.

The price to pay for such advantages is complexity. There are now two high-side and two low-side Mosfets to drive, and four output diodes. Moreover, the arrangement means that two of these Mosfets and two diodes are

in series each cycle, so the efficiency is less than ideal. If only we could get the advantages of the double-ended converter without these disadvantages!

Well, you can, by using a more complex transformer, as shown at the bottom of Fig.10.

This is the transformer-coupled half-bridge or push-pull topology, and it has all the advantages of the full bridge, but is considerably simpler. There are only two switches, and both are ground-referenced. Only two diodes are required, and the output current only ever passes through one of them. Nice.

The flyback converter

Next, I want to cover the flyback topology, which is an isolated topology derived from the boost converter (Fig.11). This time, we split the circuit in the middle of the inductor, creating two coupled sections. It looks a lot like a transformer, but strictly speaking, does not behave that way.

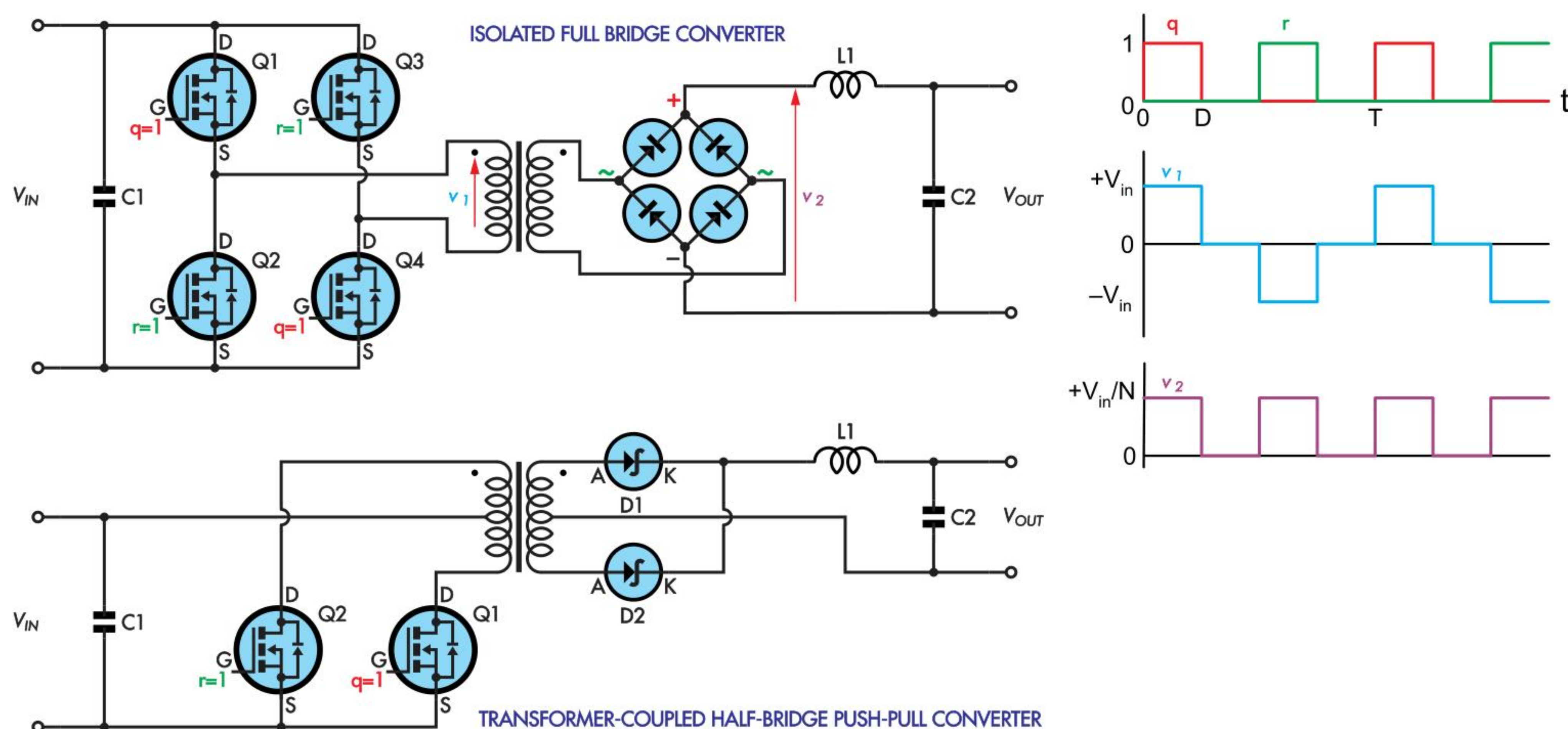


Fig.10: double-ended converters can drive flux through the transformer core in both directions, increasing the efficiency of the magnetics.

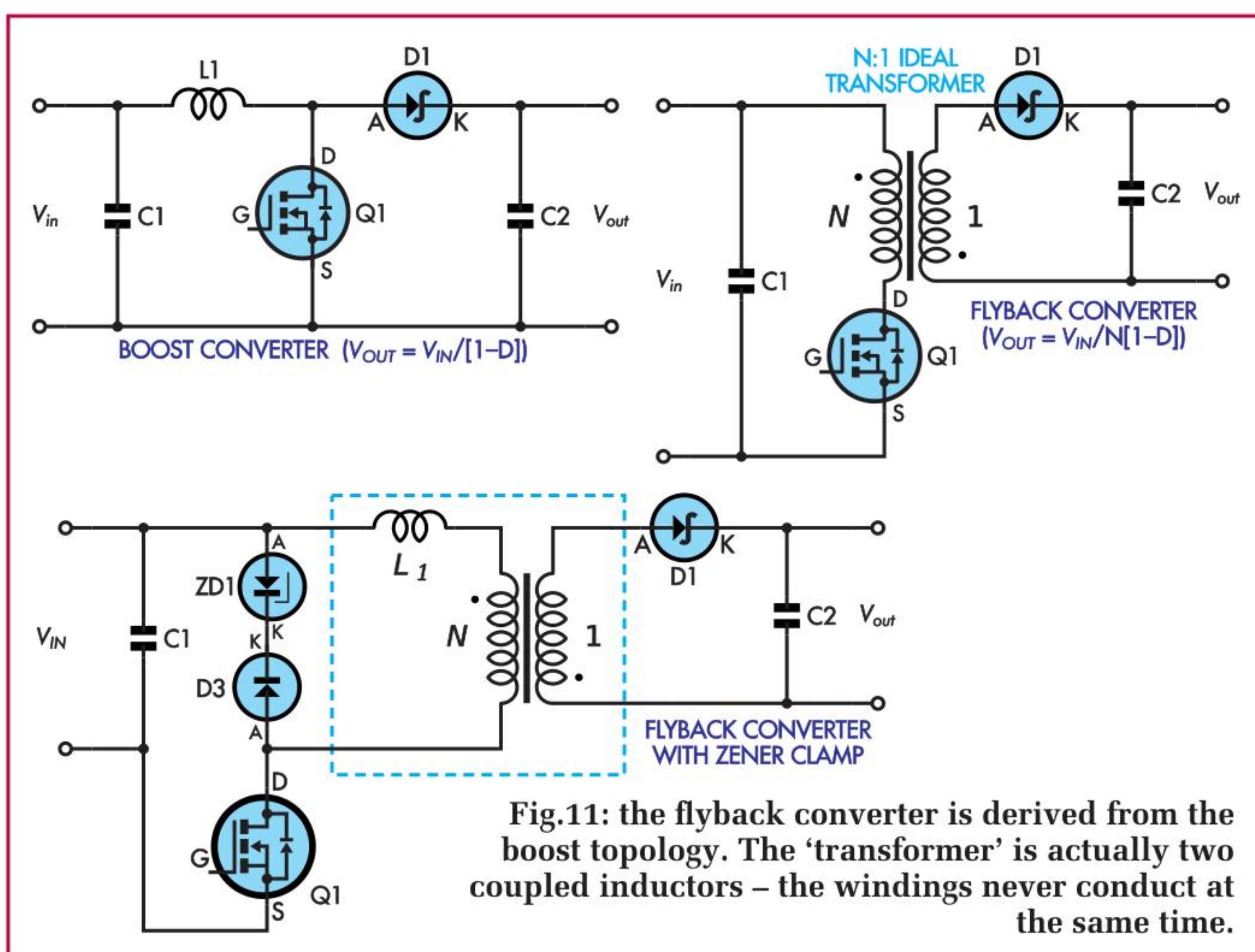


Fig.11: the flyback converter is derived from the boost topology. The 'transformer' is actually two coupled inductors – the windings never conduct at the same time.

When $Q1$ is on, current flows into the dotted primary terminal of the 'transformer'. True transformer action would require it to emerge from the secondary's dotted terminal, but it cannot, because it is blocked by the diode ($D1$). The secondary winding is effectively open circuit, so the 'transformer' acts like an inductor, building up flux and storing up energy in the core.

When the Mosfet switches off, the primary winding is open-circuited and the magnetic field begins to collapse, reversing the voltage on both windings and allowing $D1$ to conduct.

Now that the primary winding is open-circuit, the flyback transformer secondary acts like an inductor and the current ramps down, just as it does in the boost converter. The only difference is that the turns ratio means the secondary current is scaled by a factor N . The voltage transfer function for the flyback converter is therefore the same as for the boost converter, but scaled by the turns ratio.

While the flyback circuit looks a lot like the forward converter, the crucial difference is in the operation of the transformer. A forward converter has a true transformer in that the net flux largely cancels (except for the magnetising flux), and no appreciable energy is stored. The output filter inductor remains the primary energy storage element.

In flyback converters, the 'transformer' is also the energy storage element. Since the two windings never conduct simultaneously, the flux increases significantly. Flyback transformers are really two-winding inductors and usually have gapped cores. If you are ever unsure about what topology you have, take a look at the dots on the transformer, and work out if both windings can conduct at the same time or not.

The fact that flyback converters combine the energy storage element and the isolation element into one piece of magnetics (and because they

The Mornsun LM25-23B12 25W 12V isolated power supply.

use one less diode) is one of the reasons why they are the most common topology for small mains converters. That includes many phone chargers, plugpacks and other low-power DC-DC converter modules below about 50W.

Flyback converter transformers do not have the magnetising inductance concern that single-ended forward converters do (because they are not really a transformer), but they do have a problem with leakage inductance, as shown at the bottom of Fig.11.

When the Mosfet switches off, the energy stored in the core is delivered to the secondary, but any that is stored in the primary side leakage inductance has no place to go since it is, by definition, not linked by the secondary winding. So weirdly, it turns out that a practical flyback converter needs a similar type of clamp as a single-ended forward converter, but for a different reason altogether.

A professional design

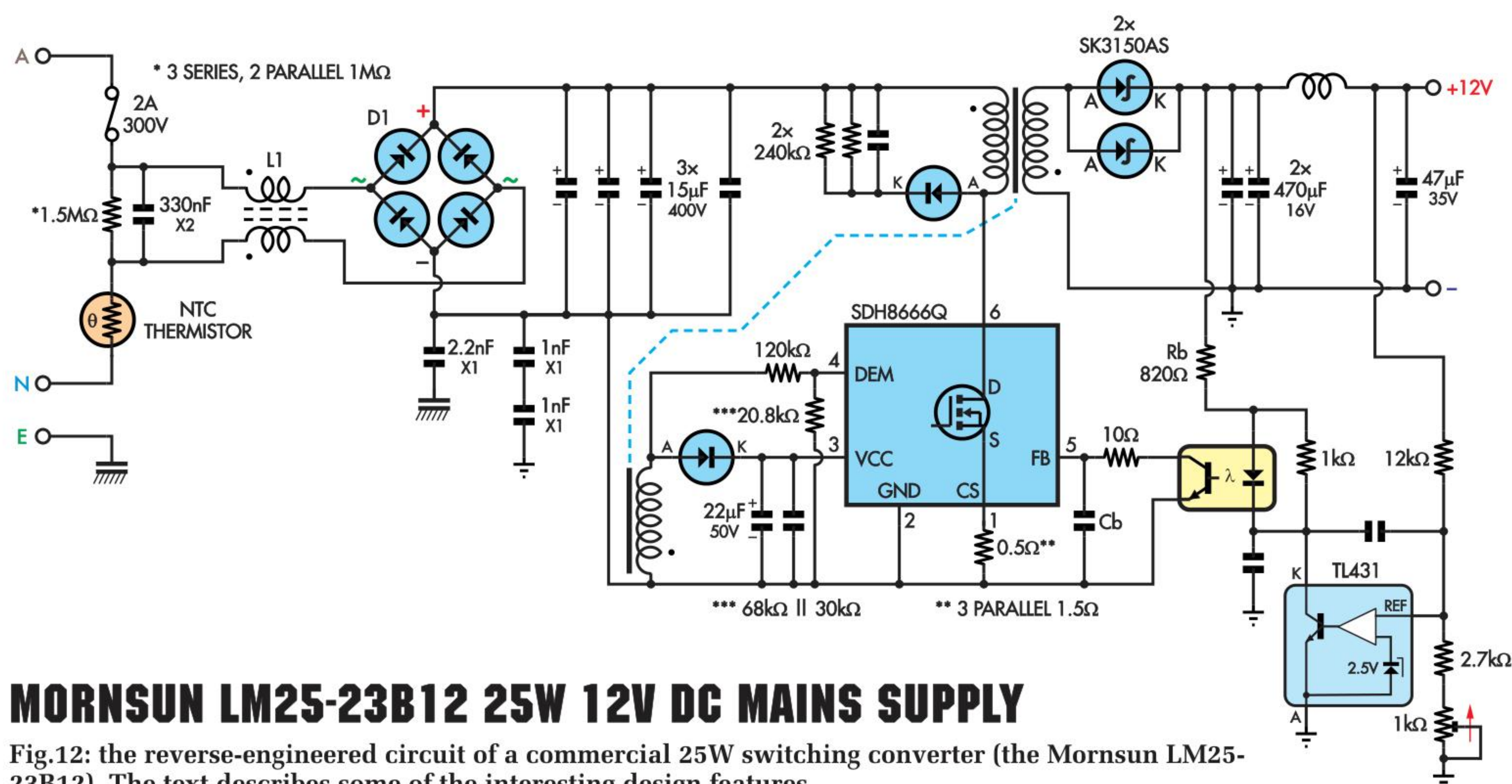
To add a practical twist, I want to take a close look at the design of a commercial flyback converter, because you can learn a lot by looking at designs by experts. The converter I chose is a Mornsun LM25-23B12, an offline isolated 25W switcher with a 12V DC output. It can accept input voltages in the range of 100V to 277V AC and can deliver 2.1A at 50°C.

You might think this is an AC-DC converter and not a DC-DC converter, and by some definitions, you would be right. Still, I will argue that, like many mains-powered supplies, it is an AC-DC converter followed by a DC-DC converter.

The AC-DC side of this converter is a simple bridge rectifier, so all the interesting stuff is happening in the DC-DC part. These converters are built to a price, but they do claim to meet a bunch of international specifications for safety & EMC (electromagnetic emissions compliance). Looking at the construction &



Part 3: Isolated DC-DC Converters



MORNSUN LM25-23B12 25W 12V DC MAINS SUPPLY

Fig.12: the reverse-engineered circuit of a commercial 25W switching converter (the Mornsun LM25-23B12). The text describes some of the interesting design features.

component choice, I don't doubt that this is a well-designed unit.

The accompanying photos show the power supply and both sides of the PCB. It is a single-sided board with through-hole components on the top side and SMT parts on the bottom. The slots milled into the board are to provide creepage isolation between primary and secondary and between high voltages.

The circuit, as best as I could reverse engineer it, is shown in Fig.12. Starting on the left is the mains input, with a fuse and an inrush-limiting NTC resistor. An X2 capacitor and common-mode inductor provide some filtering to minimise the amount of EMI (electromagnetic interference) conducted back onto the mains.

This filter is followed by a full-wave bridge rectifier and three 15μF 400V DC capacitors in parallel, to smooth the input to the flyback converter.

The negative side of the high-voltage DC supply is tied to mains Earth via a 2.2nF X1 capacitor, and to the output negative rail by two 1nF X1 capacitors in series. These capacitors provide a path to shunt high-frequency noise to Earth without compromising the safety or isolation.

The power circuit looks like any flyback circuit, with one side of the transformer primary connected to the positive supply and the other to the drain of the Mosfet switch, which is incorporated into the SDH8666Q control IC.

The Mosfet's source is connected to the current sense (CS) pin, which is connected to the negative supply via a 0.5Ω shunt resistor (three parallel 1.5Ω resistors). This chip uses current-mode control, and this is the current-sense resistor.

An RCD clamp with a 120kΩ resistor (two parallel 240kΩ resistors) and an unmarked capacitor protects the Mosfet from spikes caused by the transformer leakage inductance I described above.

On the secondary side, the rectifier consists of two SK3150AS schottky diodes connected in parallel. These are 150V 3A diodes, and I guess two are used in parallel since the peak current could easily be twice the maximum output current of 2.1A.

The diodes are followed by two parallel 470μF electrolytic filter caps. I am a bit surprised not to see a ceramic cap in parallel with these, given the switching frequency is in the 65kHz range. These caps must be working hard from a ripple current perspective (but presumably within their specifications).

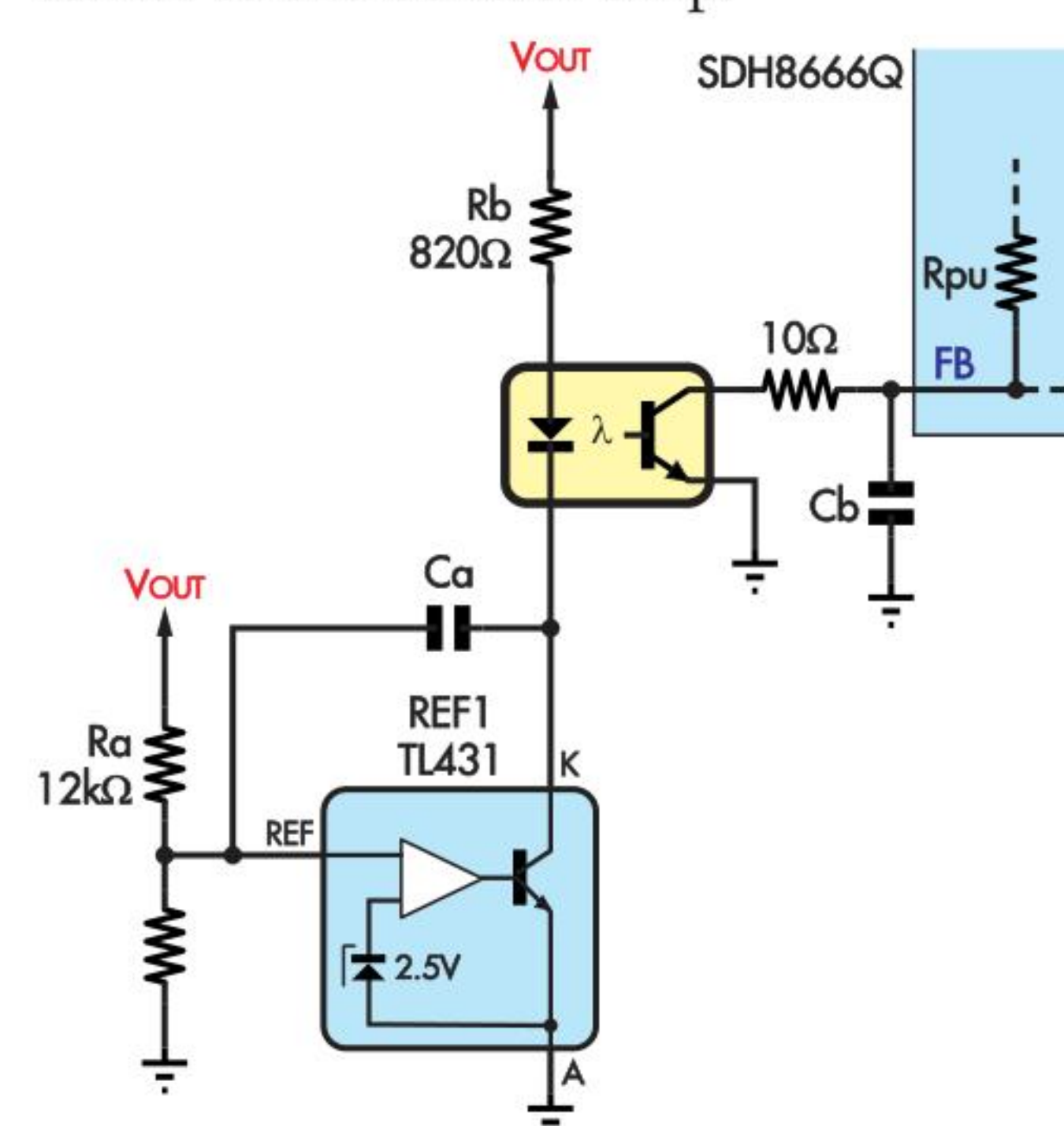
A secondary filter comprising a small inductor and a 47μF capacitor helps eliminate a lot of the switching noise on the output.

Isolated voltage feedback and control loop compensation is provided via the circuit at lower right. This uses a TL431 shunt reference and opto-coupler in a clever (but common) arrangement.

I think this circuit is worth a bit of

a closer look, so I have redrawn it in Fig.13. The converter's output voltage is divided down and compared to the 2.5V reference internal to the TL431. The resulting error voltage at the TL431's anode is converted to a current by R_b to drive the opto-coupler's LED.

A current proportional to this will flow into the coupler phototransistor's collector and be converted back to an error voltage by the pullup resistor internal to the control chip.



$$\frac{V_{FB}}{V_{OUT}} = \frac{R_{pu} \cdot CTR}{R_b} \left(\frac{1 + R_a C_a s}{R_a C_a s (1 + R_{pu} C_b s)} \right)$$

Fig.13: the Mornsun voltage feedback, error amplifier, isolation and loop compensation circuit uses a TL431, and opto-coupler and a handful of passives. The compensator is a Type II circuit, suitable for current-mode controllers, as described last month.

I calculated the circuit's small-signal transfer function using the complex impedance method we covered last month. As we would expect with a current-mode controller, the result is characteristic of a Type II compensator. I will spare you the maths.

The constant terms at the front of the transfer function relate to the opto-coupler's current transfer ratio (CTR), and the resistors on each side that convert voltage to current to voltage. The interesting part is inside the brackets.

There is a pole at the origin and another formed by the capacitor C_b and the pullup resistor R_{pu} inside the controller. There is also a zero formed by R_a and C_a . This zero cancels the output capacitance/load resistance zero, and the C_b/R_{pu} pole cancels the zero formed by the output capacitor and its ESR.

The purpose of the capacitor directly under the opto-coupler in Fig.12 is a bit of a mystery, but as it measures about 22pF and is connected at the output of the TL431's internal op amp, I suspect it is there for stability and plays no meaningful part in the control loop.

The next interesting part of the flyback converter is the power supply for the control chip. This is derived from an auxiliary winding on the flyback transformer via a diode and a couple of capacitors to supply the V_{cc} pin of the controller.

You may ask yourself how this circuit can possibly start, given that the chip is powered by its own output. The chip has a clever trick up its sleeve in that there is an internal high-voltage depletion-mode (normally on) Mosfet connected to the DRAIN pin (pin 6) that initially provides a small trickle current to charge the 22 μ F capacitor on the V_{cc} pin.

To keep the power dissipation in the depletion Mosfet to a minimum, this current is very small – nowhere near enough to power the chip – so the chip is not enabled until the V_{cc} voltage reaches some fairly high predetermined level. At this point, the chip switches on and uses the charge stored in the 22 μ F cap to run for long enough for the external supply to take over.

Once everything is up and running, the depletion-mode Mosfet is switched off to save energy.

This leaves only the DEM pin, which

is fed from the auxiliary secondary winding prior to the rectifier diode. This pin is used to (roughly) sense the output voltage to provide output overvoltage protection, and something called "valley lockout", which appears to be a mechanism to prevent the chip restarting too quickly due to small dips in the input voltage.

The DEM pin can indirectly sense the converter's output voltage because the voltage on the auxiliary winding is proportional to the output voltage (less one diode drop) according to the turns ratio.

I suspect that the valley lockout works by ignoring short dips in the input voltage (sensed at the DRAIN pin) if the output voltage (sensed at the DEM pin) does not also drop. This would prevent the start-up sequence described above from happening unnecessarily for very short mains interruptions that don't impact the output.

That's it for this month. Next month, we will have a look at AC-to-DC converters, and we will go through the design process for a simple DC power source in detail. We will build and test the circuit to see how well the theory and practice align. **PE**

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2SA970BL **£1.50**
2SC2204, 2SD655, BC550C **£0.50**
ZTX651 **£0.30**
ZTX751 **£0.50**

RF transistors (suitable for Theremin)
BF199 **£0.50**

Audio power MOSFETs
Exicon 10N20, 10P20 **£6.50**
Hitachi 2SJ99, 2SK343 **£3.50**
Hitachi 2SJ56, 2SK176 **£8.50**

JFETs
BFW11, BFW10, TIS73L, J177, J113,
U1994, U1898, 2SJ176, J201 **£1.00**
J175, J176, J112, J111, 2N3820,
2N5457, BF244, 2N5460, J230 **£0.60**

Small power output/driver transistors
2SB649A, D669A, 2SA1208,
2SC2910, MJE253G, MJE243G,
2SA1725, 2SC4511 **£1.20**
BD139, BD140, BD135, BD136, BD435,
BD436, MPSA63, BCV46 **£0.50**

MOSFETs
ZVP2106A **£0.30**
Dual-gate 3SK45, BFS28 **£1.50**

Metal-cased transistors
BCY71 **£0.30**

Amplifiers
LM318 high-speed op amp **£0.35**
µA709 metal-cased op amp **£2.00**
5534H metal-cased op amp **£3.00**
LM384, TDA2030A, TDA2050V **£1.50**
TAA435 (Mullard power amp
driver for AD161/AD162) **£1.20**

Synthesiser ICs

That/dbx 2180 VCA/VCF **£6.00**
CA3080 VCA/VCF (vocoder) **£2.80**

CA3280 VCA/VCF **£4.00**
LM13600/ LM13700 VCA/VCF **£2.00**
CA3086 transistor array **£1.00**
PT2399 echo/delay **£1.50**

Dual transistors

2N2639, 2N2223, 2N2910 (NPN) **£4.00**
E401 (JFET Moog) **£4.00**
2SK2145-Y dual JFET **£0.80**
2N5564 JFET **£8.00**
HN3C51F, HN3A51F **£1.00**
DMMT3904/6, HN1A01F,
HN1C01F **£0.50**

Loudspeakers

PE Mini-Monitor
Volt PE165 6.5-inch woofer (each) **£85**
Morel MDT29 tweeter (each) **£25**
Kit pair of PE165/MDT29 plus
Volt crossover parts and PCBs **£299**
Monacor DT-28N tweeter (each) **£35**
Vifa 19mm BC20SC15-04
tweeter (each) **£15**

Volt crossover inductors
1.2mH, 1.5mH, 2mH, 2.7mH, 0.5mH
(tapped at 0.3mH) (each) **£5.85**

*Fully assembled and tested high-quality
speaker prototypes – ask for details*
LS3/5As and other similar speaker
systems (pair) **£200-£350**

Fully tested reclaimed speakers
Vifa BC14 5-inch woofer (each) **£10**
Vifa TC26 1-inch tweeter (each) **£10**

Low-price speakers
Philips 4-inch 4070 **£2.00**
EMI 10x6-inch, 30Ω Alnico **£7**
64mm 64Ω neodymium **£1.20**

Capacitors

Note '10/63' denotes '10µF 63V'.

Polyester
3.3/100, 4.7/250, 4.7/63 **£1.00**
5.6/63, 8.2/63, 10/63 **£2.00**
Mullard 'Mustard' C296 0.22/400 **£2.00**

Polycarbonate
Axial 2.2/63 1%, 4.7/160, 6.8/63 **£1.00**

Radial
6.8/160V, 10/63 **£2.00**
Reclaimed 22/63 **£2.00**

Polystyrene
Philips 1% 4.7nF/160, 6.2nF/500,
12nF/63, 22nF/63, 110nF/63,
24nF, 2nF **£1.00**
RIFA 1% 100nF/100, Suflex
90.9nF 0.5% **£2.00**
Suflex 2.5% 10nF/63 (rad. or ax.) **£0.50**

Radiation resistant
Siemens cellulose acetate MKL
2.2/25 **£0.80**

Electrolytic – Mullard blue 017 series
10/25, 22/25, 100/10 **£0.50**
150/40, 470/40, 1000/40 **£1.00**

Tantalum – axial metal cased
22/50, 47/35, 68/25, 100/20, 120/10,
150/16, 220/10, 330/6 **£2.00**
22/35, 33/35, 47/20, 68/15, 100/10,
150/6 **£1.25**
4.7/50, 6.8/35, 10/25, 10/35, 22/15 **£1.00**

Axial moulded-case tantalum
Kemet axial 6.8/10 **£0.30**
Kemet radial 33/10 **£0.40**
STC radial 100/20 **£1.50**

Tantalum bead
22/50 **£1.00**
470/3 **£2.00**
680/6.3 **£3.00**

Wet tantalum
220/25 axial **£3.20**
Castanet button 140/30, 470/3 **£3.20**
Hughes 540/10 **£3.20**

Bipolar
Hermetic bipolar tantalum 16/35 **£3.20**
Elcap axial 10/50 **£0.50**
Generic radial 100/16, 470/35,
100/ 63, 22/35, 4.7/35, 220/16 **£0.50**

Philips solid-aluminium (axial) 121/123
47/16 **£1.00**
330/6.3 **£2.00**
100/35 **£4.00**

Philips Pearl 122 series (radial)
10/16 **£0.30**

Silvered mica (radial)
1nF/500 1% **£1.00**

Trimmer capacitors
Vishay plastic-film 4-40pF **£0.80**
Vishay plastic-film 5.5-45pF **£1.00**
Vishay plastic-film 5-80pF **£1.20**
Vishay plastic-film 10-250pF **£2.00**
Mica 1-12pF, 2-40pF **£1.00**

Audio transformers and inductors

Eagle transformers
LT44, LT722 driver, LT700, LT723
500Ω output **£2.50**
LT30 500mW output **£3.50**
5:1 interstage (fuzz box) **£1.50**

Repanco
T/T3 splitter transformer **£4.00**
CH2 5mH RFC **£2.00**

Balanced output transformer
Vigortronix 600Ω VTX-101-007 **£10**
Vigortronix 600Ω VTX-101-3001 **£10**
Vigortronix 600Ω VTX-101-3002 **£15**
Gardners 150Ω **£10**
Reclaimed BBC LL74/MPC
nickel core 600Ω **£12**
Low-level 1:1 600Ω (modem) £2
Reclaimed mic input transformer **£15**

Inductors
82µH, 4.7mH, 100µH, 270µH,
10µH, 14µH (low Z) **£0.50**
4-inch ferrite rod with MW
and LW windings **£3.00**

Special resistors

Bourns wire-wound trimmer
5kΩ 3059 JM panel-mount **£2.00**
Thermistor RA53, R13 **£4.00**
A13 **£2.00**
Thermistor CZ1, CZ6 **£1.50**
Holco H2 2.2MΩ 1W, 1% **£1.00**
Welwyn 1GΩ 2W **£1.00**
5k Bourns 3321H cermet trimmer **£0.50**

Potentiometers

Bourns 81
25kΩ lin cermet **£2.00**
25kΩ lin, 5kΩ lin cond. plastic **£3.00**

Bourns 91
10k dual-gang lin or log **£10.00**

Plessey moulded-track
5kΩ log with switch **£3.00**
50kΩ A/log **£3.00**
Mil 250kΩ lin dual **£5.00**

Alpha
16mm 4.7kΩ A/log **£0.80**
220kΩ A/log **£0.80**
10kΩ lin centre-detent dual-gang **£1.50**

Allen Bradley J series/Honeywell
10kΩ lin 1 million cycles **£5.00**
Blore Edwards AB 45 dual 5kΩ
A/log with switch **£3.50**
Alps RK9 dual-gang 5kΩ RD law **£4.00**
Alps 50k log tapped motor. stereo **£5.00**
BI P260 500kΩ log conductive
plastic 1 million cycles **£2.00**
Colvern wire-wound 100kΩ or
50kΩ dual-gang 3W **£5.00**
Mil Spec hermetic 10Ω **£8.00**

Miscellaneous

PCBs
All-pass Filter PCBs (three) **£4.00**
.. Rotary Switch (AO0626ROT) **£1.50**
.. Pot Board (AO0626POT) **£1.50**
Silent Stomp Switch (pair) **£4.00**
Penta-fuzz Bender (PE, Feb26) **£3.00**
Pre-driver (PE, Jan26) **£3.00**
Pre-driver jack bd (PE, Jan26) **£1.50**
Fuzz Box (PE, Nov25) **£3.00**
Fuzz Box jack bd (PE, Dec25) **£1.50**
Dual JFET Buffer (PE, Oct25) **£2.00**
Mosfet Buffer (PE, Oct 2025) **£2.00**
Dual Mosfet Buffer (PE, Oct25) **£2.00**
Cascode Amp (PE, Sep 2025) **£2.00**
Cascode PS (PE, Sep 2025) **£2.00**
Pocket Theremin (EPE, 1996) **£2.00**
Elysian Theremin (EPE, 1996) **£6.00**
Synth VCF, VCO (EPE, 2017) **£3.00**
48V PSU (EPE, 2019) **£3.00**
SMT dual transistor adapter **£0.40**

Relays
STC hermetic 4190 24V DPDT **£4.00**

Switches Blore Edwards/Plessey
MH1 4-pole 3-way **£3.00**
Elma 01 2-pole 6-way **£6.00**
3PDT guitar pedal stomp switch **£2.50**

Valves
EC88, PCL84 **£3.00**

Contact

Jake Rothman
Manchester House
Market Street, Craven Arms
Shropshire SY7 9NN
(visit by appointment)
+44 (0)1597 829102
jrothman1962@gmail.com

Minimum order £5.00 inc post

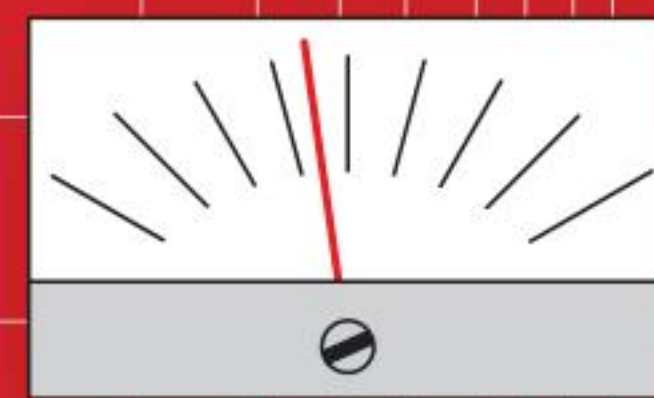
Payment

PayPal, bank transfer, cheques
(payable to "J Rothman", UK
pounds only). No VAT payable

Postage

Small Jiffy bag £2.99
Small package £4.99
Big boxes and overseas at cost
– ask for a quote

AUDIO OUT



By Jake Rothman

All-pass filters, part three – building an all-pass filter

We have been looking at the implementation and uses of all-pass filters over the last couple of issues. After a bit more details on their use in loudspeaker crossovers, we'll shortly get to a useful one to eight stage all-pass filter board you can build from standard parts.

Funny filters

KEF used asymmetrical filters with second-order bass and third-order treble slopes for many of their designs. This provided some extra filter delay for the tweeter, compensating for its physical offset. They then finished the job by moving the tweeter up the baffle a bit.

This technique was executed perfectly on their 104aB speaker (Photo 17) and no all-pass filter was required. The tweeter crossover filter circuit shown in Fig.30 for the T27 drive unit gave a proper third-order

acoustic Butterworth slope with correct phase, hence the aB suffix.

The T27's high-Q fundamental resonance became inaudible. The new network improved on the original network so much that you could buy a drop-in replacement crossover to upgrade it.

The circuit effectively removed the bump from the tweeter's resonance and removed its additional second-order slope from the roll-off. The difference was clearly noticeable on massed strings, the most revealing test for crossover problems. This filter has been used on many systems, such as the KEF 101, the Wilmslow Audio CS1a and Malcolm Jones's Falcon LS3/5A derivatives.

Also see my article in the September 2019 issue of *PE* on building LS3/5A style speakers.

I use the same circuit topology in most of my designs as well. The KEF

104aB has a lovely laid-back sound, but this was really a defect, resulting from the lack of off-axis midrange energy. Using a 200mm bass unit up to 3.5kHz is too high and results in a very directional output.

This then crosses over to a wide directivity output as the tweeter comes in. This results in an inaccurate 'dog-leg' response curve off-axis.

The LS3/5a had a complex phase equalisation arrangement of an asymmetrical wide overlap crossover and a tweeter with polarity inversion. The flanges for the drivers almost overlapped. To accommodate all these variables, they mounted the bass unit on the rear of the front baffle to introduce additional displacement (see Photo 18).

The result was almost perfect integration between the two drivers. There's a little resonance due to the extra cavity introduced in front of the



Photo 17: the KEF 104aB speaker is one of my favourites, apart from the auxiliary bass radiator loading (not shown), which gives bad bass group delay. The upgraded crossover is on the small removable panel. I replaced the fuse with a PTC using the hole for a tweeter level control.

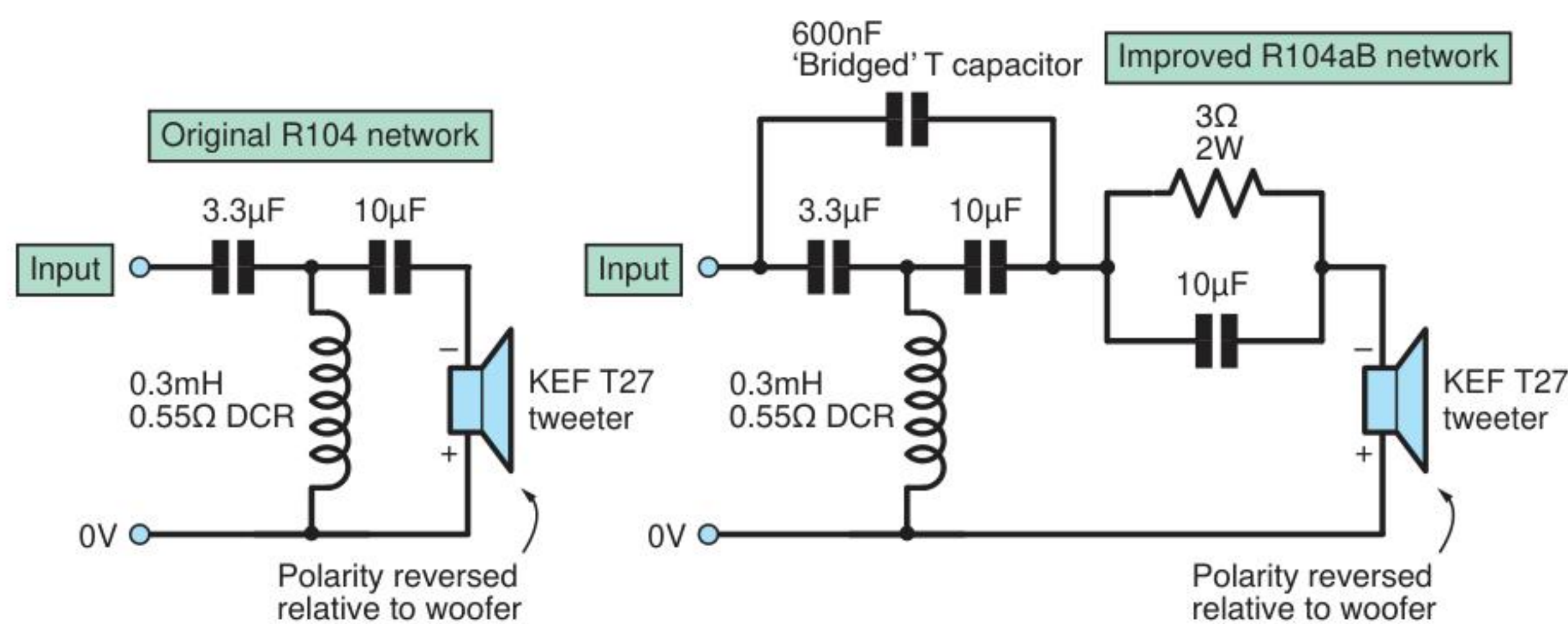


Fig.30: KEF pioneered the use of computers to design crossover networks that had a total acoustic output following the correct curve. In this case... it's a third-order Butterworth filter.

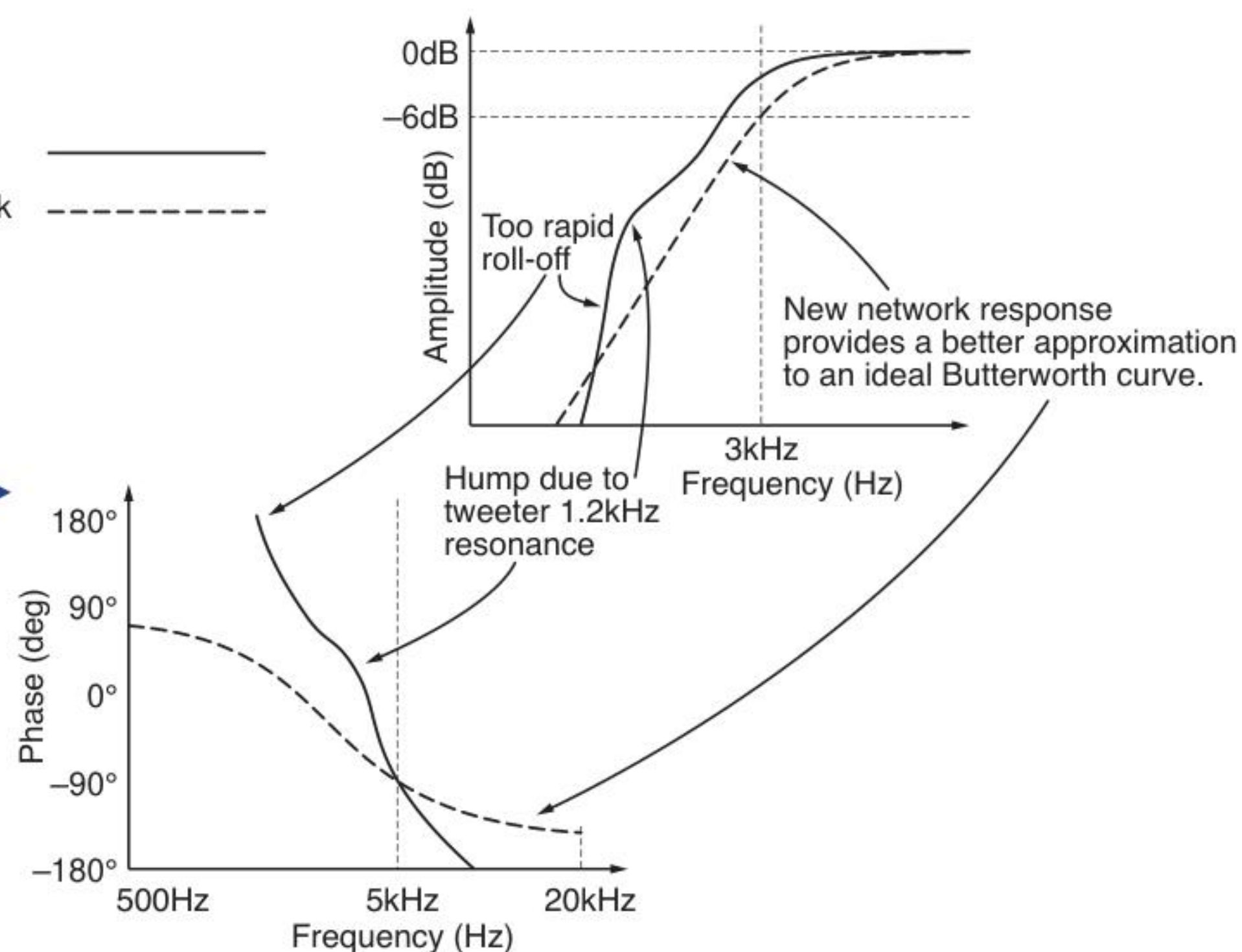




Photo 18: the famous BBC LS3/5a has the woofer mounted on the rear of the baffle for extra delay.

cone, which I don't like. When I make "LS3/5As", I instead front-mount the bass driver and use an all-pass filter.

Subtractive filters

One interesting point about all-pass filters is that the phase shift is double that of a high-pass or low-pass filter of the same order because the phase shift effect of one capacitor is 90° in a frequency filter and 180° in an all-pass filter. This is useful to know if you want to make subtractive filters.

An example of this would be trying to make a low-pass filter by subtracting a high-pass filtered signal from the direct signal. If this is done as shown in Fig.31, it doesn't work because the phase shift introduced by the filter messes up the

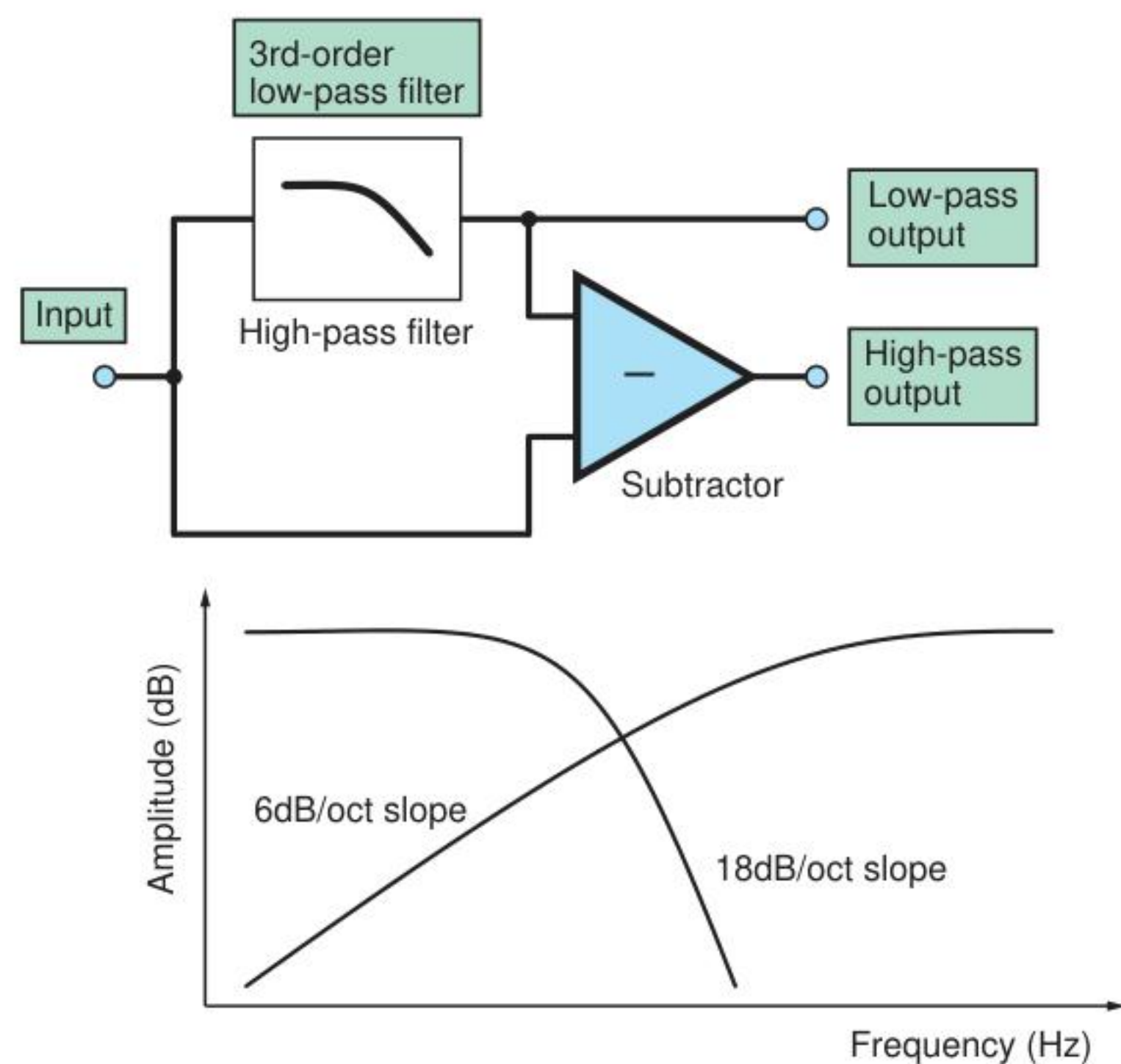


Fig.31: if the output of a low-pass filter is subtracted from the direct signal, a mirror-image high-pass filter should result, but it doesn't work properly because the phase shifts upset the subtraction process.

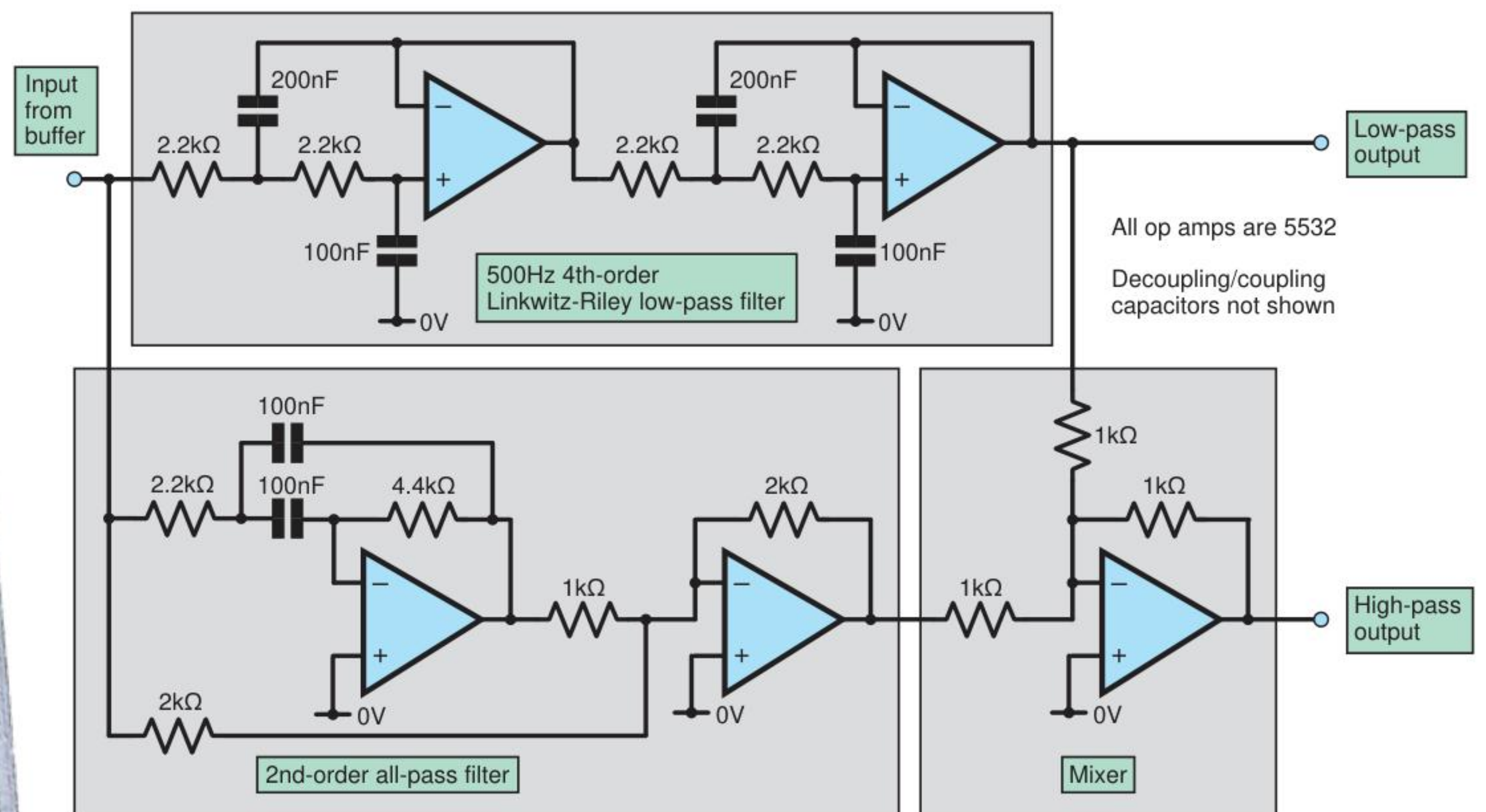


Fig.33: a 500Hz fourth-order linear-phase subtractive crossover filter. Notice the second-order all-pass network.

subtraction, resulting in a first-order slope even if the high-pass filter is of a higher order.

Adding an all-pass filter is the way to make it work, as shown in Fig.32. This gives you the added advantage that the overall filter responses, when combined, have a constant group delay, resulting in what is incorrectly called a linear-phase crossover filter. This has the advantage of reducing off-axis notches in the speaker's response. It also makes the bass-to-mid crossover more time-coherent. However, all subtractive systems are dependent

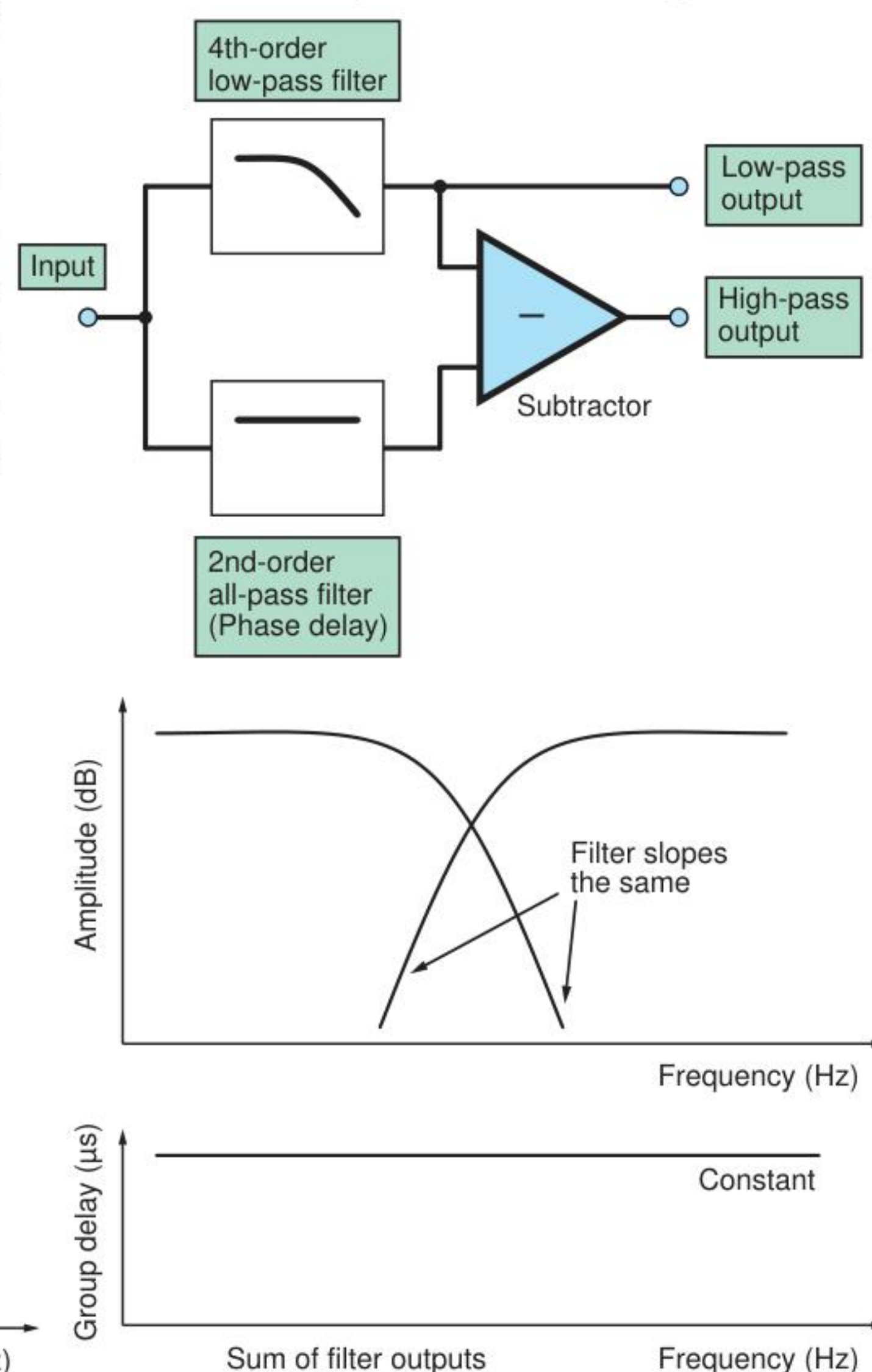


Fig.32: the subtraction process can be corrected by adding a phase-compensating all-pass filter. The crossover filter becomes 'linear-phase', an audio marketing term for constant group delay with frequency.

on accurate component values to ensure complete cancellation.

A residual bass signal being fed into a tweeter could damage it and cause intermodulation distortion. It may be possible to trim this out, but it would involve not only amplitude but phase correction as well. I could just imagine the response of a production engineer if I presented a three-way active speaker with eight trim adjustments!

Having said that, it is not a problem for home construction because tweaking to perfection is part of the fun. I could specify high-precision $\pm 1\%$ polystyrene capacitors, but then the accountant would have a heart attack. Linn use these for their studio monitors, but at £15k a pair, I would expect the best components!

A bit of residual bass would do no harm on the low/mid crossover, so this would be the best place to try a linear-phase crossover, especially as it's the low end, where varying group delay is most audible. Fig.33 shows a 500Hz fourth-order version. I fancy designing a PCB for this.

The advantage of a phase-linear crossover is that the on-axis lobe can be wider before the nulls occur. This is because there is a less rapid phase shift around the edges of the crossover frequency region. These phase-linear crossover filters were developed by Stanley Lipshitz and John Vanderkooy and should not be confused with the Linkwitz-Riley (LR) crossover (I know I have!).

The LR crossover provides good phase-related directivity control, but has a peak in group delay around the crossover region and a dip in the idealised overall acoustic power response. Importantly, that does not necessarily mean there will be a dip in the perceived acoustic response.

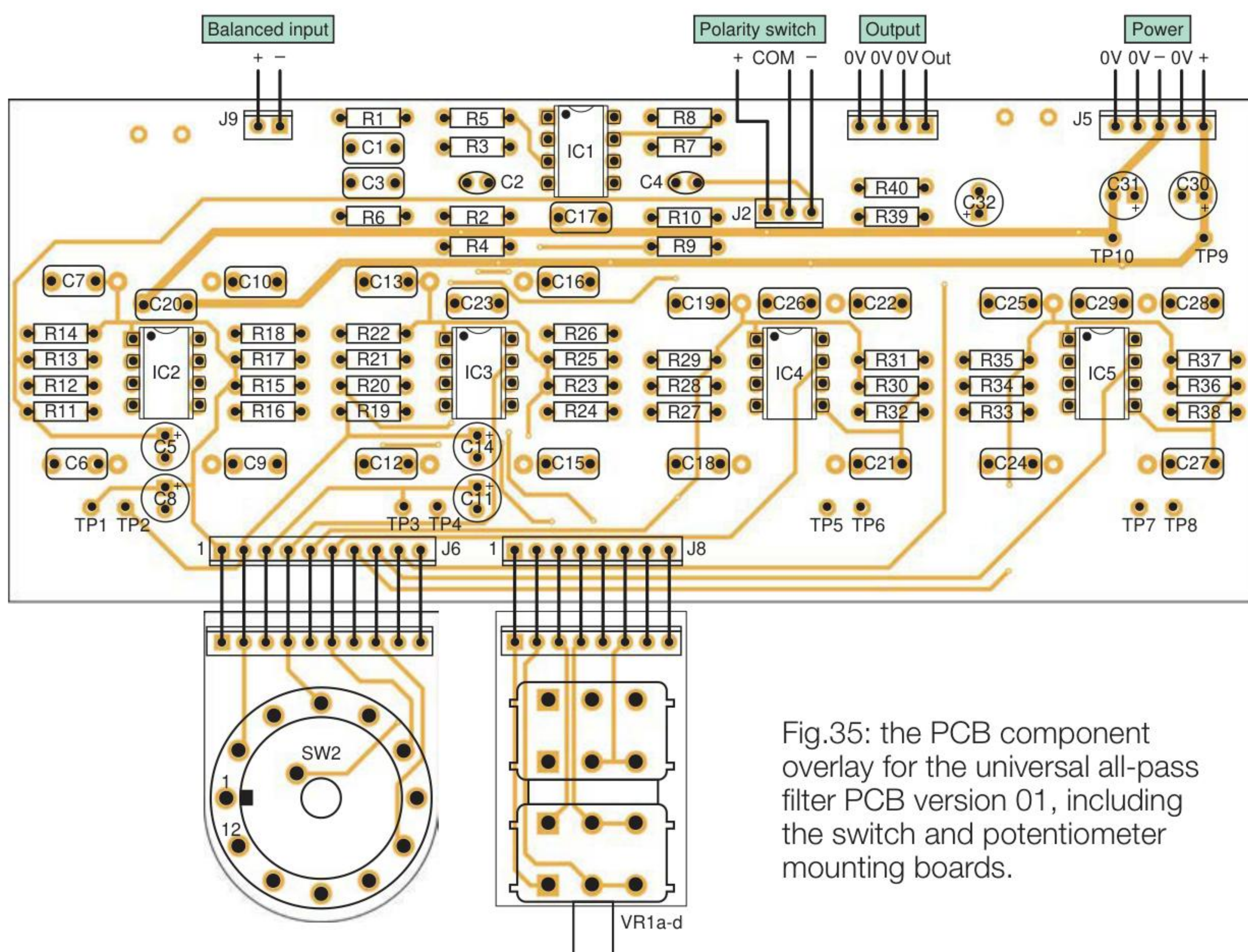


Fig.35: the PCB component overlay for the universal all-pass filter PCB version 01, including the switch and potentiometer mounting boards.

To expand on that last point, an LR crossover is designed so that each output is 6dB down at the crossover frequency. On-axis, the acoustic output of the loudspeaker at this point is the result of constructive interference between the two drivers, so the pressures simply sum.

Since the signals sent to the two drivers are identical at the crossover frequency, the two acoustic outputs add coherently, resulting in a doubling of pressure (+6dB). The -6dB level from each driver therefore sums to give a flat on-axis response. Without this -6dB level in each branch, the two drivers would produce an acoustic peak at the crossover frequency.

The off-axis behaviour is more complex. There is usually a large dip in the vertical off-axis response, depending on the driver spacing and directivity.

Soldering time

Enough theory for now; it's time to get building. I suggest using the generic op amp boards described in the December 2022, January 2023 and February 2023 issues to build the switched inverter, attenuverter and single or two stage all-pass filters. For the more complex circuits, the universal all-pass filter circuit shown in Fig.34 is needed.

This has up to eight cascaded all-pass sections or poles available, built around four dual op amps. Note that 16 stages (two units in series) are required for the effect to be clearly

heard on sounds such as kick drums, where the thump will appear to lag behind the click.

Grindle has designed a board; its overlay diagram is shown in Fig.35. There is a rotary switch to select the number of all-pass sections required (SW2) and a four-gang potentiometer (VR1) to provide tuning. It would be nice to vary all eight sections to get the biggest adjustment range possible, but I've only recently found eight-gang pots as stock items (eg, the Bourns PTD908-1015F-B103 from Mouser). Those could be used in a future design.

Both these controls can be omitted to make a fixed delay unit if desired. The provision of the pot connector (J8) allows a modulator board to be plugged in to make a sweepable studio phaser. This could use LM13700 transconductance op amps, VCAs, Vactrols or optoisolated JFETs. Ordinary JFETs won't work here, since they have to be referred to ground; these are better used for a 9V guitar phaser pedal, which we'll describe later on.

There is a balanced input section to drive the all-pass filter chain, with a polarity switch, SW1. This is built around two differential stages wired in antiparallel to give equal input impedances on both input pins.

There seemed no point in employ-

ing a dedicated single op amp (an NE5534) here. Surprisingly, the dual version is cheaper than the single version, and every audio engineer has tubes of 5532s.

Poka-yoke

The connectors on the PCB are all different sizes, in accordance with the Toyota poka-yoke assembly technique, to avoid mistakes. I blew a board up recently because the output and power supply connectors were both three-pin Molex plugs. I'm never going to make that mistake again!

The completed PCB is shown in Photo 19.

The switch and pot are on little adaptor boards shown in Photos 20 & 21. Both of these

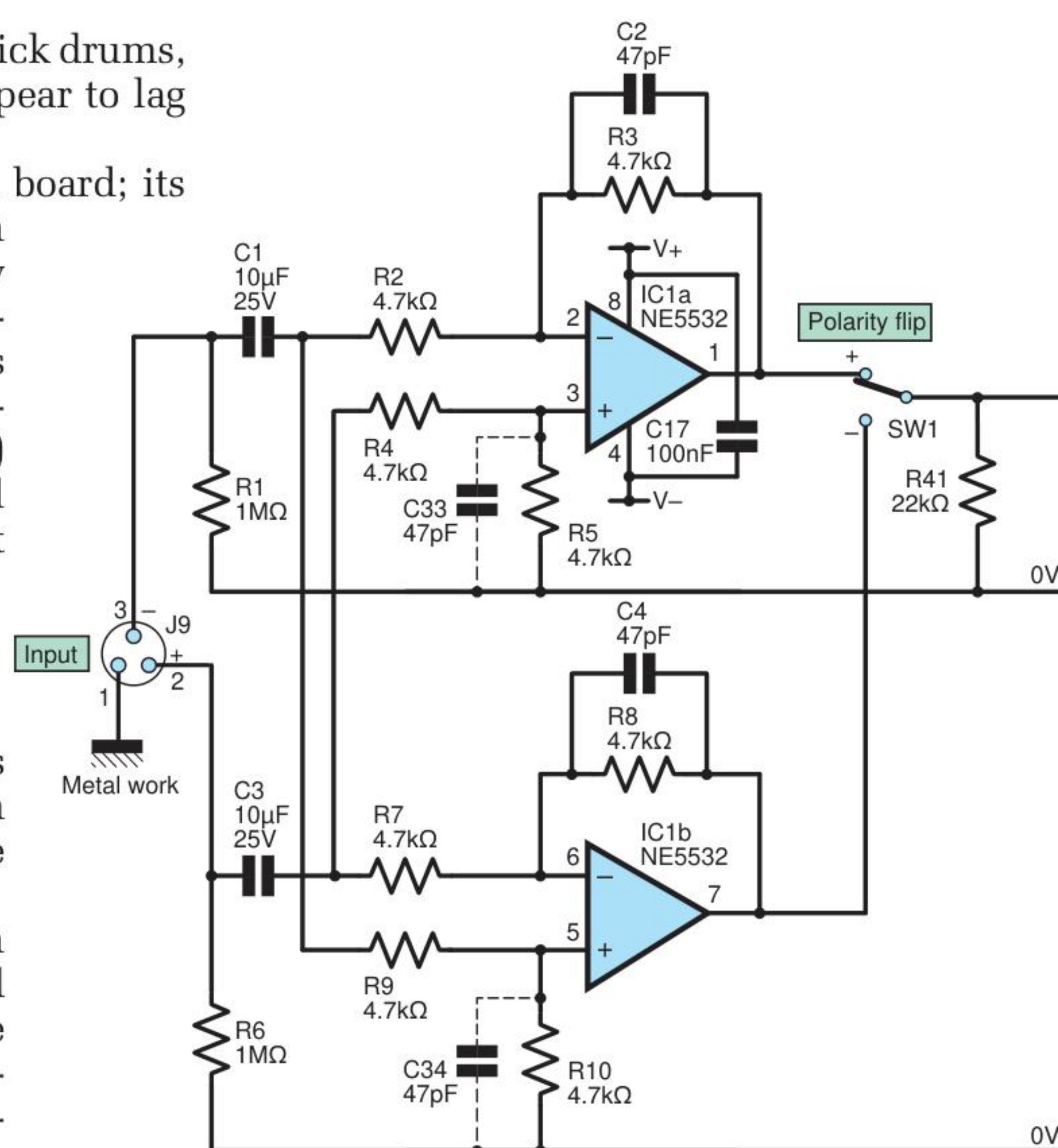
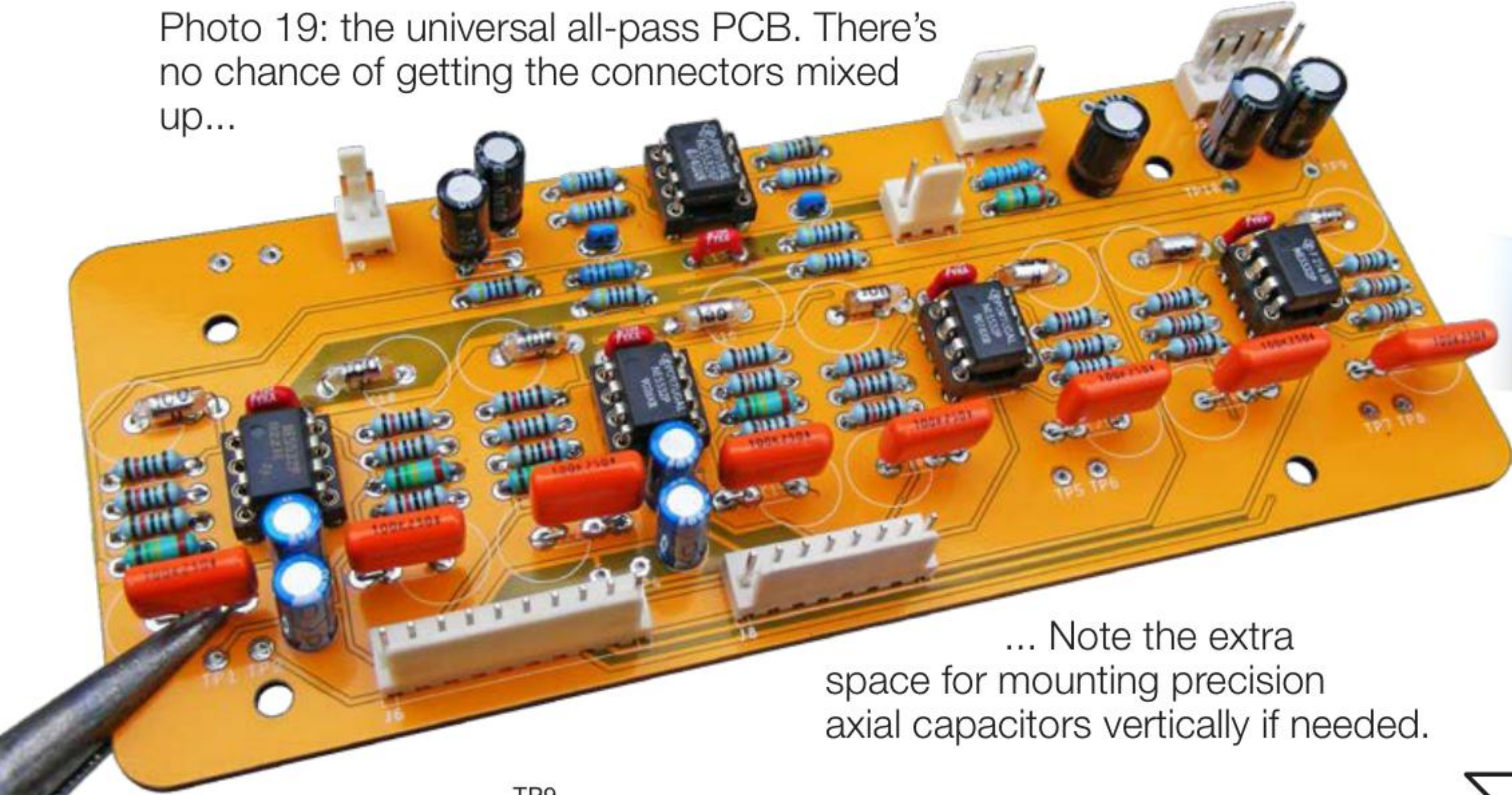


Photo 21: the quad-gang pot also has its own PCB. This could be useful for other filter projects.

Photo 19: the universal all-pass PCB. There's no chance of getting the connectors mixed up...



... Note the extra space for mounting precision axial capacitors vertically if needed.

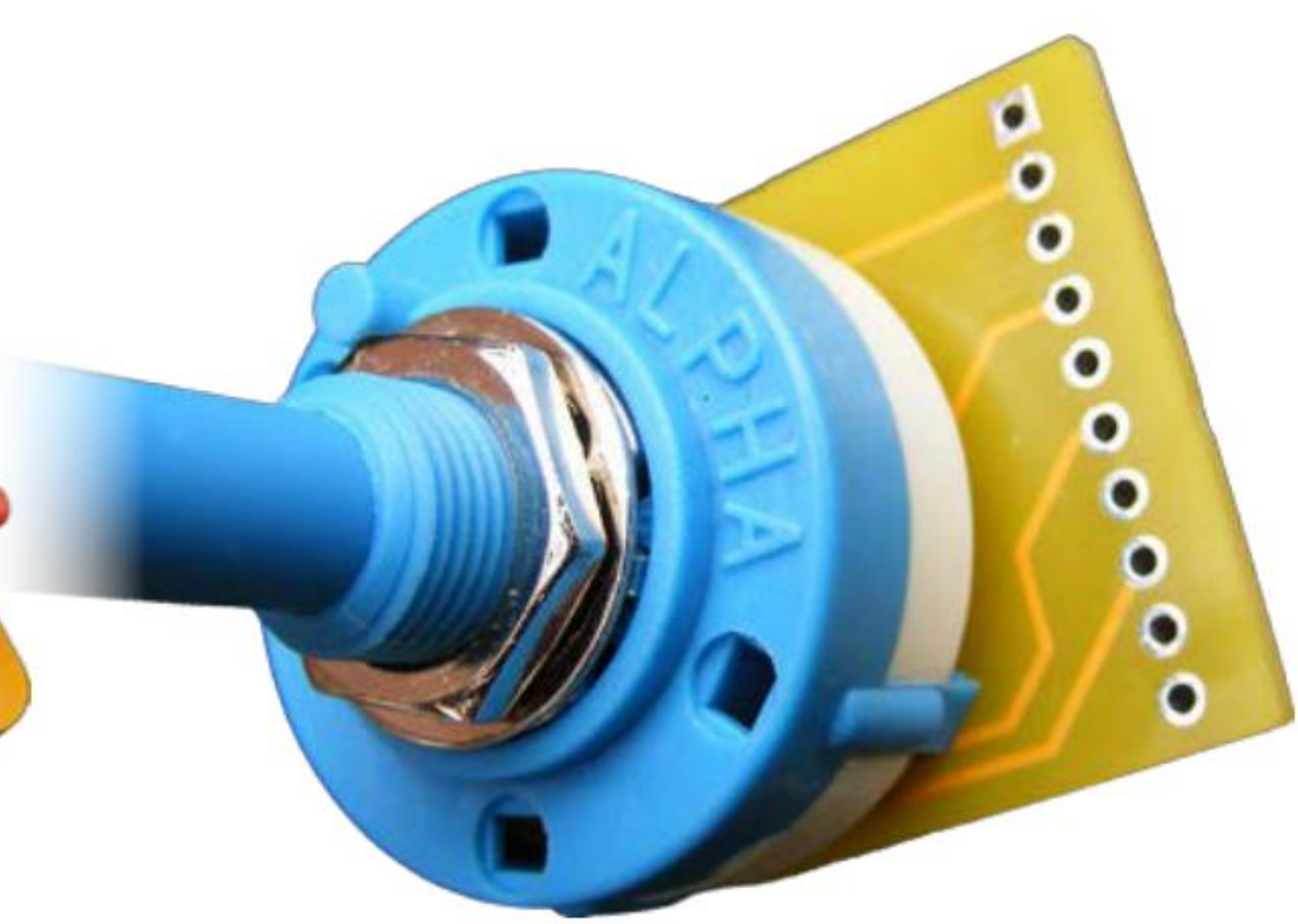
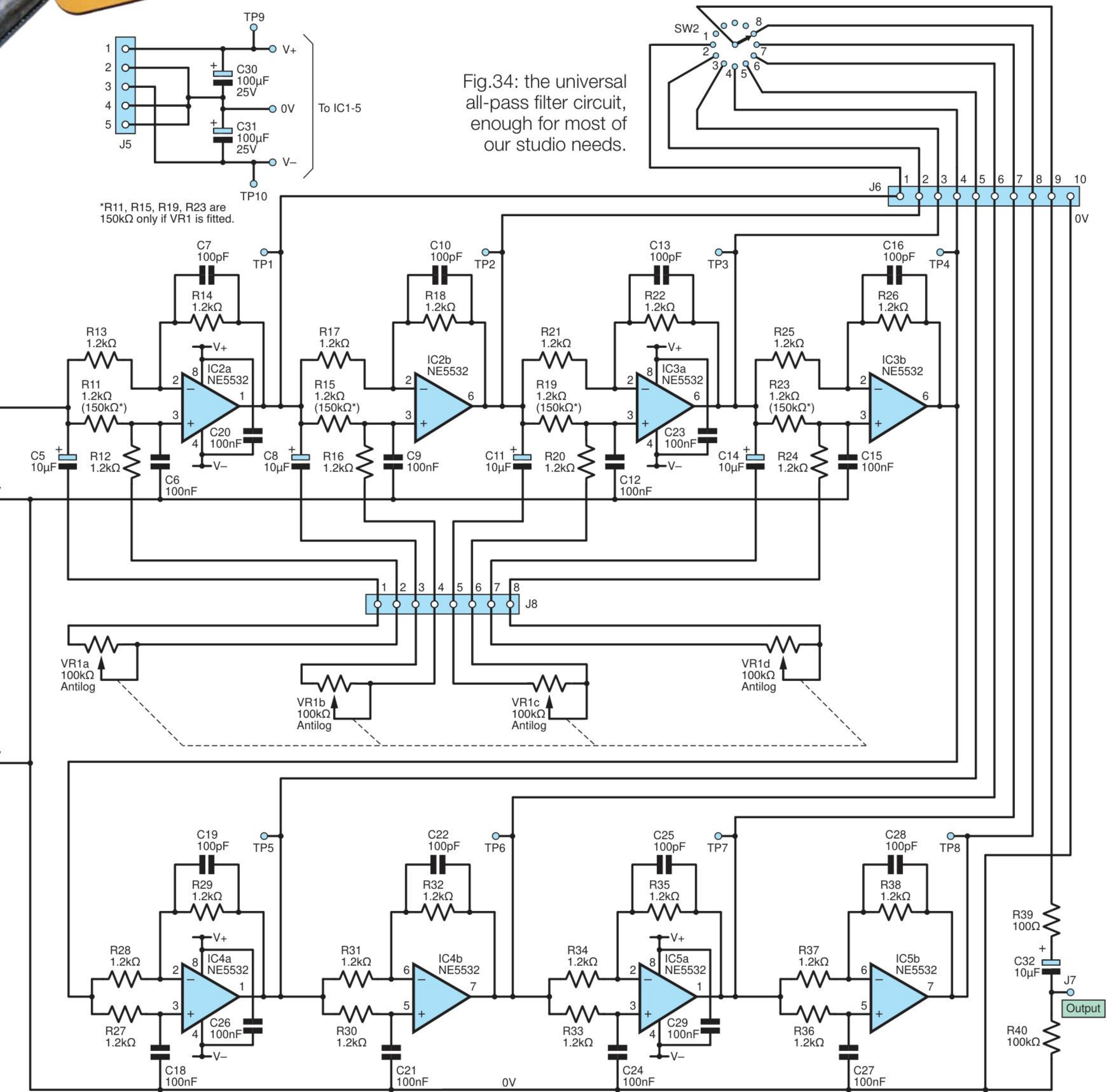


Photo 20: the switch is mounted on a sub-board with a handy SIL header connection.



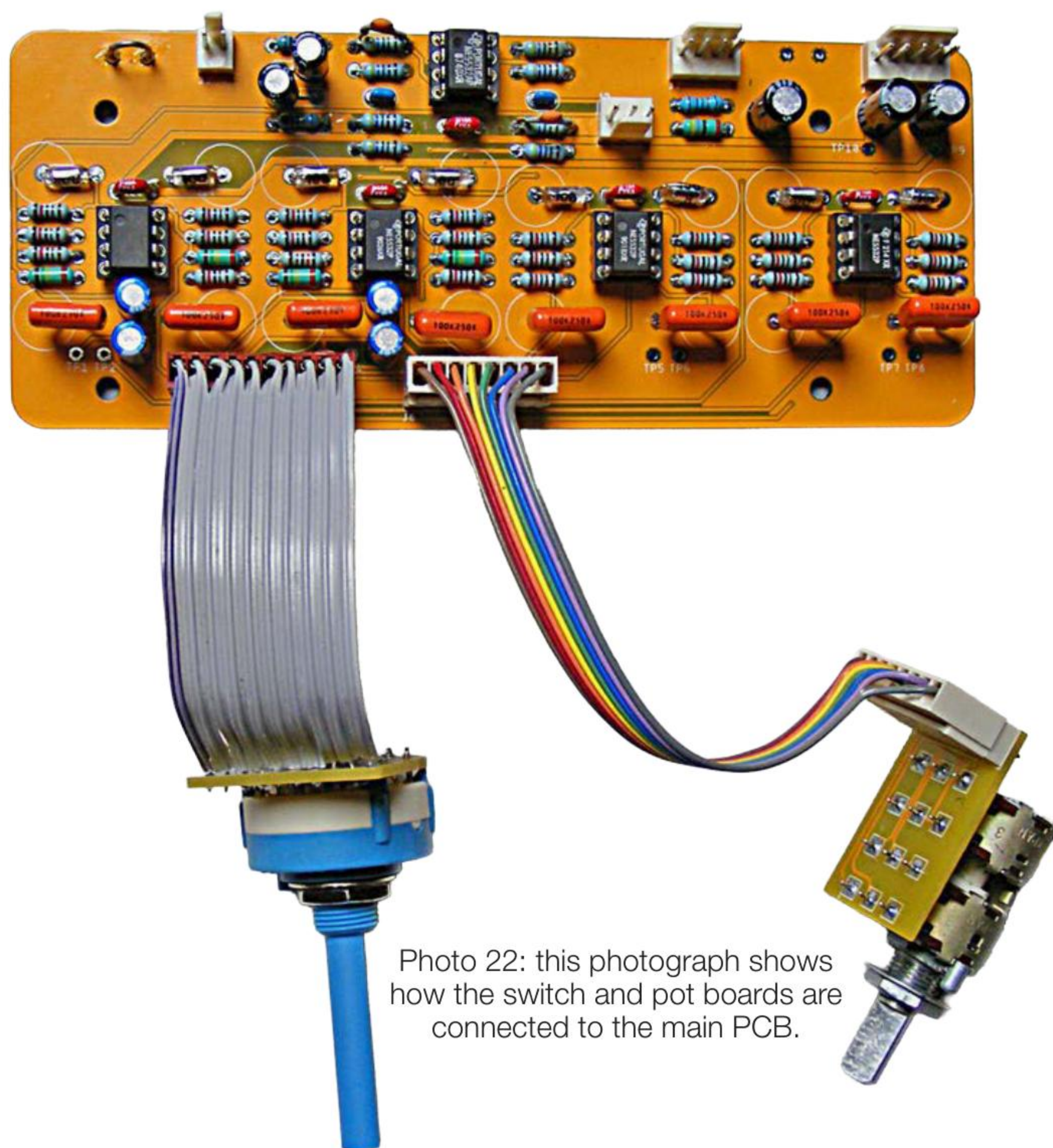


Photo 22: this photograph shows how the switch and pot boards are connected to the main PCB.

boards could be handy in other projects. They put all the wiring in a neat line so that ribbon cables can be used. This is shown in Photo 22; it is much quicker to assemble than the tangle of different lead lengths that wiring such tagged components normally entails.

These boards are so useful in their own right that they are available separately. The pot and switch are not mounted on the main board, since this gives uneven heights on the control panel and the components would be subject to more mechanical strain.

A 12-way single-pole rotary switch from Alpha or Lorlin is used because of its standard pinout. Pin 1 is marked on the board, and this should enable the other pins and the wiper pin to all line up to

push the board on with a bit of judicious bending of the tags. Remember to install the rotation limit washer at position number eight, as shown in Photo 23.

Fitting the components

All the resistors associated with all-pass stages are 1.2k Ω because this is the lowest value that can be accommodated, giving the lowest noise. 1.2k Ω is the limit because the lowest impedance the NE5532 standard low-noise audio op amp can drive without significantly increased distortion is 600 Ω . The two input resistors, R11 and R13, are effectively in parallel at high frequencies.

If the tuning pot is to be omitted, resistors R11, R15, R19 and R23 have to be selected to give the desired frequency (from 1.2k Ω up to 100k Ω) as well. If the pot is used, the resistors have to be increased to 150k Ω to maintain the DC path.

Minor modifications

To avoid a loud click when polarity switch SW1 is operated, it's a good idea to fit a 22k Ω pull-down resistor from the input to the first all-pass filter to ground (R41). I inadvertently forgot this component when giving the design brief to Grindle! This can easily be fitted

under the board from the centre pin of the switch connector to the ground end of R40.

Two other small parts I forgot are capacitors C33 and C34 on the input balancing amps for RF filtering. Simply solder them across R5 and R10. These mods are shown in Photos 24 & 25. I'll be doing a mounting board for SW1, which will incorporate R41 and a few other resistors and capacitors to give minimum clicking.

Powering it

The board should be run on a dual-rail supply from $\pm 12\text{V}$ up to $\pm 21\text{V}$ (the absolute maximum for the NE5532 is $\pm 22\text{V}$). The noise does not increase as the voltage increases, so running at the highest voltage possible gives higher headroom and therefore a better signal-to-noise ratio (given sufficient signal). However, the power consumption and dissipation increase proportionally.

The quiescent current consumption per chip is 8mA, so with five chips on the board, that's 40mA. Given a total power rail of 44V, that's a steady power consumption/heat dissipation of 1.7W. In the worst case, loading at full power (clipping at 10kHz) would be 20V into 600 Ω , giving 33mA per op amp. This could possibly add up to a shocking 330mA or 6.6W.

I've never known this to happen in practice, since the op amps shut down, but it's worth bearing in mind when dimensioning the power supply. Using a standard LM317/LM337 regulated supply with decent heat-sinks, such as the one I published in the April 2022 issue, will do the job for up to four boards.

That's sufficient detail for the audio alchemists among you to get started. Next time, I'll provide some specific applications for the all-pass board, such as a speech symmetriser, giving all the component values, wiring details and tests.

PE

Photo 24: minor modifications; there are always some! Capacitors are added across resistors R5 and R10.

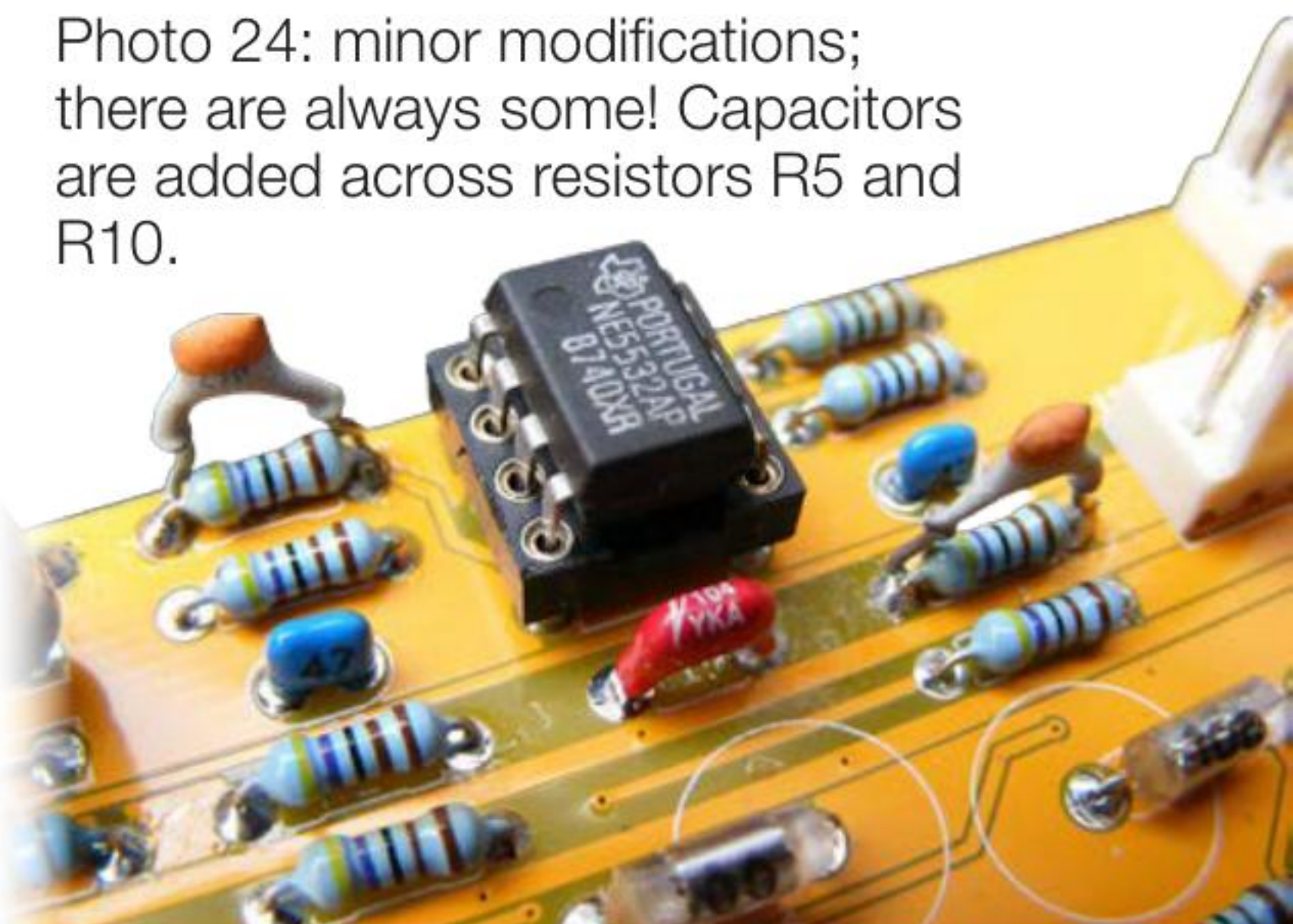


Photo 23: setting the switch to 8 positions with the special washer.



Parts List – 8-Way All-Pass Filter Board

- 1 double-sided filter PCB coded A00626ALL, 144 × 58.5mm
- 1 double-sided rotary switch PCB coded A00626RSW, 27 × 36mm
- 1 double-sided potentiometer PCB coded A00626POT, 22 × 34mm
- 5 NE5532 dual low-noise, low-distortion op amps, DIP-8 (IC1-IC5)
- 1 100kΩ quad-gang anti-log (C law) 16mm potentiometer (VR1)
[Alps, Alpha, Tayda, AO shop or Blore Edwards]
- 1 small SPDT toggle switch (SW1)
- 1 single-pole, 12-way PCB-mounting break-before-make rotary switch (SW2)
[Rapid or Tayda]
- 1 3-pin 2.54mm pitch polarised header with matching plugs and pins (J2)
- 1 5-pin 2.54mm pitch polarised header with matching plugs and pins (J5)
- 1 10-pin 2.54mm pitch polarised header with matching plugs and pins (J6)
- 1 4-pin 2.54mm pitch polarised header with matching plugs and pins (J7)
- 1 8-pin 2.54mm pitch polarised header with matching plugs and pins (J8)
- 1 2-pin 2.54mm pitch polarised header with matching plugs and pins (J9)
- 5 8-pin DIL IC sockets (for IC1-IC5)

Capacitors (all ±20% tolerance or better)

- 3 100µF 25V radial electrolytic (C30, C31 & C32)
- 4 10µF 25V radial electrolytic (C5, C8, C11 & C14)
- 2 10µF 25V radial non-polarised electrolytic (C1 & C3)
- 8 100nF* 63V polyester/MKT (C6, C9, C12, C15, C18, C21, C24 & C27)
- 5 100nF 50V X7R ceramic, 5mm pitch (C17, C20, C23, C26 & C29)
- 8 100pF 50V C0G/NP0 ceramic, 5mm pitch (C7, C10, C13, C16, C19, C22, C25 & C28)
- 4 47pF 50V C0G/NP0 ceramic, 2.5mm pitch (C2, C4, C33 & C34)

* value and dielectric depend on the application; this is a good starting point

Resistors (all ¼W ±1% axial metal film)

- 2 1MΩ (R1 & R6)
- 4 150kΩ (R11, R15, R19 & R23) (only required if VR1 is fitted)
- 1 100kΩ (R40)
- 1 22kΩ (R41)
- 8 4.7kΩ (R2-R5 & R7-R10)
- 28 1.2kΩ (R11-R38; R11, R15, R19 and R23 not required if VR1 is fitted)
- 1 100Ω (R39)

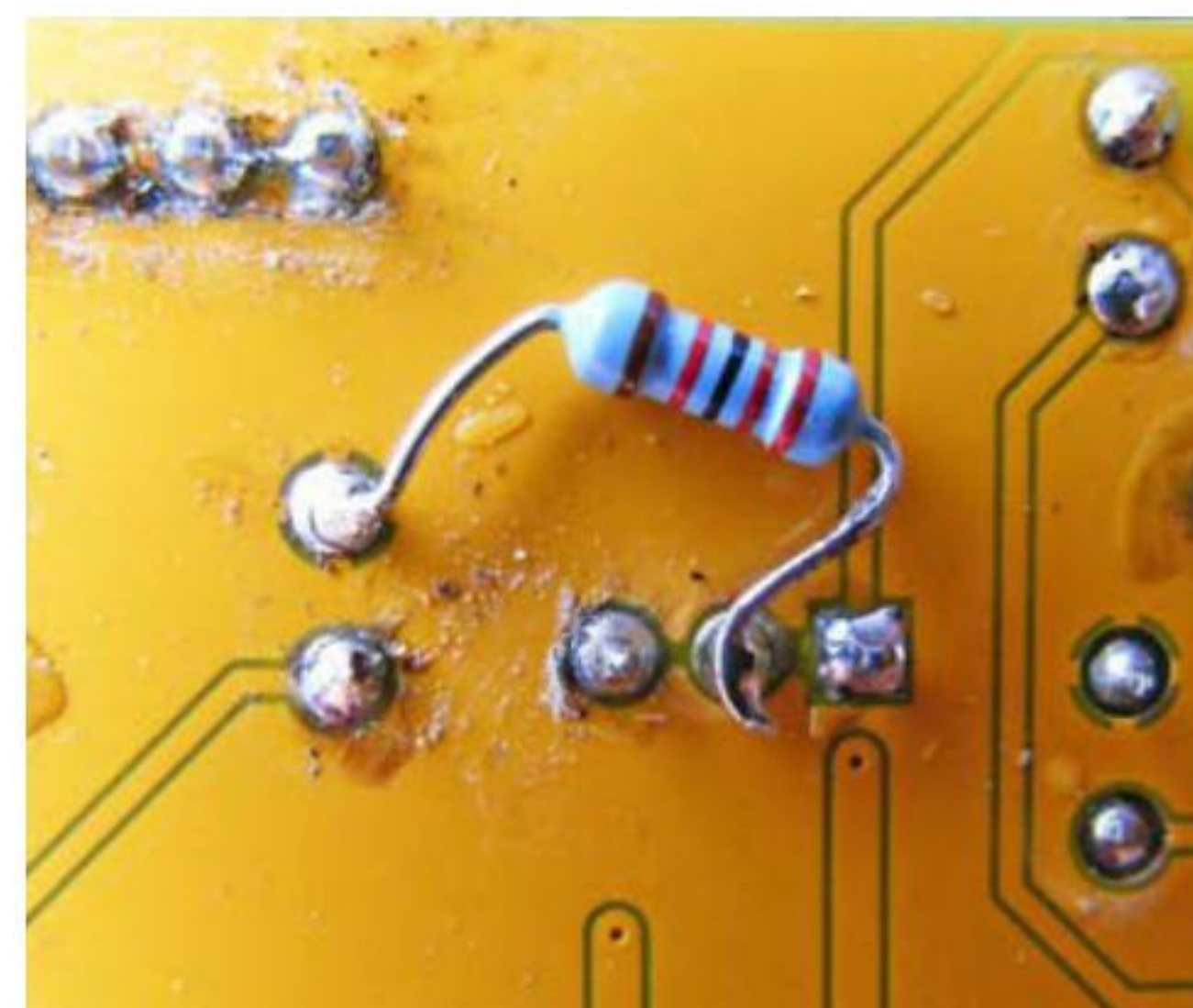


Photo 25: another minor change. An additional DC path resistor, R41, is mounted under the board from the centre terminal of the polarity switch SW1 to the ground end of R40.

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From battery-powered AI to nuclear-fuelled data centres, the race for connectivity and intelligence is becoming a race for energy efficiency.

If there's an elephant in the room or a fly in the soup (I never metaphor I didn't like), it's that it's becoming challenging to write a technology-related column these days without AI popping up in one form or another to bid us a cheery "hello". Since I don't like challenges (it's not in my vocabulary), I'm not even going to try to hold back the tide.

Don't forget the battery!

How much power things consume is important, especially in remote and portable devices, where the internet 'rubber' meets the real-world 'road'. Things get a little hairy when we talk about AI-enabled, battery-powered wearables like smartwatches and augmented reality (AR) glasses.

It's only getting worse with features like conversational user interfaces that let users interact with devices via spoken natural-language commands.

This all reminds me of an old joke about a somewhat absent-minded professor type character staggering down a busy city street, lugging an enormous and clearly very heavy suitcase. A trio of well-heeled young financiers observe him approaching, and their leader decides to have a bit of fun.

As the old fellow draws level, the leader calls out, "Hey granddad, do you have the time?"

With much grunting and wheezing, the man lowered his suitcase and pulled back his jacket sleeve to reveal a space-age contraption strapped to his wrist. Upon a spoken command, a high-resolution, full-colour 3D hologram of the Earth springs into existence, complete with real-time clouds and other atmospheric effects, slowly rotating and displaying the time across every time zone.

The snotty young men stop laughing and gawk, mouths open in wide-eyed astonishment. "I must have that," says the leader at last. "I'll write you a million-pound cheque here and now!"

The professor pauses for a moment, then shrugs. "Well, that would certainly fund my other experiments... all right."

Cheque and watch duly exchanged, the three young men set off down the street. "Just a moment!" calls the professor after them, gesturing at the suitcase.

"Don't forget the battery!"

How low can you go?

The reason the wristwatch example springs to mind is that I was just chatting with the folks at Ambiq (www.ambiq.com). Their claim to fame is ultra-low-power processors that enable edge AI in billions of battery-powered devices—everything from wearables and smart glasses to smart cameras and industrial sensors.

The lower the supply voltage, the lower the power consumption. Most state-of-the-art application processors in devices like smartphones and wearables typically operate in the 0.7–1.2V range. Ambiq dares to be different with its subthreshold power-optimised technology (SPOT).

Ambiq's forthcoming Atomiq device will boast an Arm Cortex-M55 CPU and an Arm Ethos-U85 NPU core dedicated to AI acceleration, both running with core voltages of just 0.3V. As far as I'm concerned, this is skirting the boundary between engineering and black magic!

What makes this particularly interesting is the class of applications Atomiq unlocks: always-active devices that don't need to be tethered to the cloud or constantly recharged.

Think conversational AR glasses that can respond to spoken commands all day, health-monitoring wearables that never sleep, battery-powered cameras performing real-time object recognition, or industrial sensors quietly applying AI at the edge without anyone having to worry about where the nearest power outlet is.

Feeling connected

Not long ago, apart from telephones that were firmly tethered to the wall, things largely operated in splendid isolation. Early embedded systems did their jobs quietly and efficiently, but they did them alone, with no expectation of communicating with anything else. A thermostat controlled the heating, a smoke detector monitored the house; that was pretty much the end of the story.

Today, we increasingly expect everything to be connected all the time. Using technologies like Wi-Fi, Bluetooth, Thread and Zigbee, devices are expected to discover each other, communicate and collaborate seamlessly.

Once everything is connected, the next question becomes: what do we do with all that data? This is where companies like Synaptics (www.synaptics.com) come into play. Synaptics explores ways to make devices more intuitive. That involves embedding intelligence into everyday interfaces and quietly reimagining how humans engage with machines and data.

Synaptics describes the SYN765x as a Wi-Fi 7 AI-native connected microcontroller, a device designed from the ground up to combine high-performance wireless connectivity with on-device AI processing. That makes it perfect for smart hubs, security cameras with local object recognition, or even industrial nodes making decisions without waiting for the cloud.

From local links to global reach

Connectivity doesn't stop at Bluetooth and Wi-Fi. Today's smartphones



seamlessly hop between networks, typically preferring Wi-Fi when it's available and falling back to the mobile network when it's not. However, there are still plenty of places where terrestrial coverage is patchy at best: remote regions, oceans, mountains and the occasional stubborn 'not spot' in the middle of a city. This is where low Earth orbit (LEO) satellite constellations enter the fray, extending connectivity far beyond the reach of traditional infrastructure.

It won't be long before access to these non-terrestrial networks (NTNs) becomes commonplace. Future devices are likely to treat connectivity as a hierarchy: try Wi-Fi first, then the mobile network, and finally satellite if all else fails.

The result is a world in which we (and our devices) are effectively never out of touch, with a continuous thread of communication stretching from the living room to the most remote corners of the planet.

At present, terrestrial 5G networks and satellite-based systems are still somewhat different beasts, each with its own standards, constraints and challenges. CEVA (www.ceva-ip.com) recently announced its PentaG-NTN platform, which is designed to enable seamless integration between mobile and satellite communications.

By providing the building blocks for multi-mode devices that can operate across both terrestrial and non-terrestrial networks, companies like CEVA are helping to turn the vision of truly ubiquitous connectivity into reality.

Powering the cloud

Power consumption is becoming a major concern in the cloud and in data centres, where the numbers can get eye-wateringly large very quickly.

Training AI is the process of building the model in the first place—feeding it vast quantities of data and adjusting millions or billions of parameters until it behaves as desired. This is

computationally intensive and can consume enormous amounts of energy, often requiring thousands of GPUs running for days or even weeks.

After the model has been trained, while the system is actually used—recognising images, understanding speech, generating text and so on—the energy usage is far lower each time. However, this usage is multiplied by the number of inferences made, which can be in the millions or even billions.

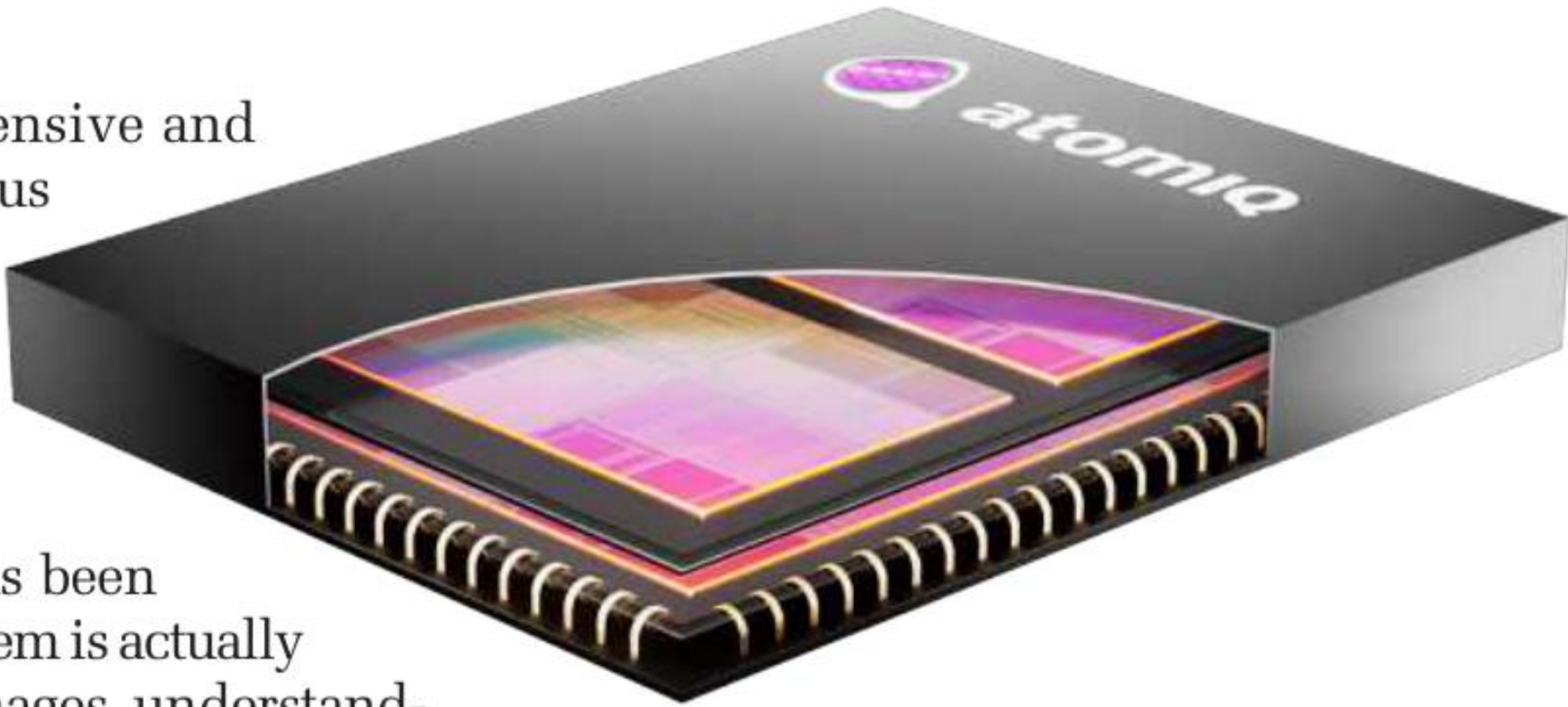
So while training represents a large, one-off (or at least occasional) energy investment, making inferences is where the real long-term power consumption lies. Every time someone asks a question, issues a voice command, or triggers an AI-enabled device, more power is used. And that adds up.

The scale of this demand is already reshaping the infrastructure that supports it. Some companies are even exploring small modular reactors (SMRs)—essentially compact nuclear power stations—to keep their AI data centres fed. When nuclear-powered data centres start to sound like a sensible option, you know things have escalated rather dramatically.

How much?

I designed my first application-specific integrated circuit (ASIC) back in the dark ages we used to call 1980. This was at the five-micron node, meaning the smallest structures we could create on the surface of the silicon die (chip) were 5 millionths of a meter across.

Today, people routinely create chips at the five-nanometre node (and below), which means transistors are around 1000 times smaller and consume correspondingly less power. Well, sort of... The feature sizes aren't quite as small as 5nm might suggest, but they are still very tiny.



ASICs in the 1980s contained only a few thousand transistors (if you were lucky). By comparison, a high-end AI processor today can contain tens to hundreds of billions of transistors. So, while each transistor may be far more efficient than its ancestors, we now use vastly more of them.

Modern AI accelerators can each draw 500-1000 watts. Pack eight of these into a server and you're looking at several kilowatts per box. Fill a rack with such systems, and you can easily reach 20–50 kilowatts per rack. Multiply that by tens of thousands of racks in a hyperscale data centre, and you begin to see why power consumption has become such a pressing concern.

When less is more

Until recently, I would have said that a custom ASIC offers the highest possible performance while consuming the least possible power. That's still true as far as it goes, but there's a wrinkle. A custom ASIC accelerator, like a GPU, is designed to handle a wide variety of workloads, including AI training and inference, as well as other tasks.

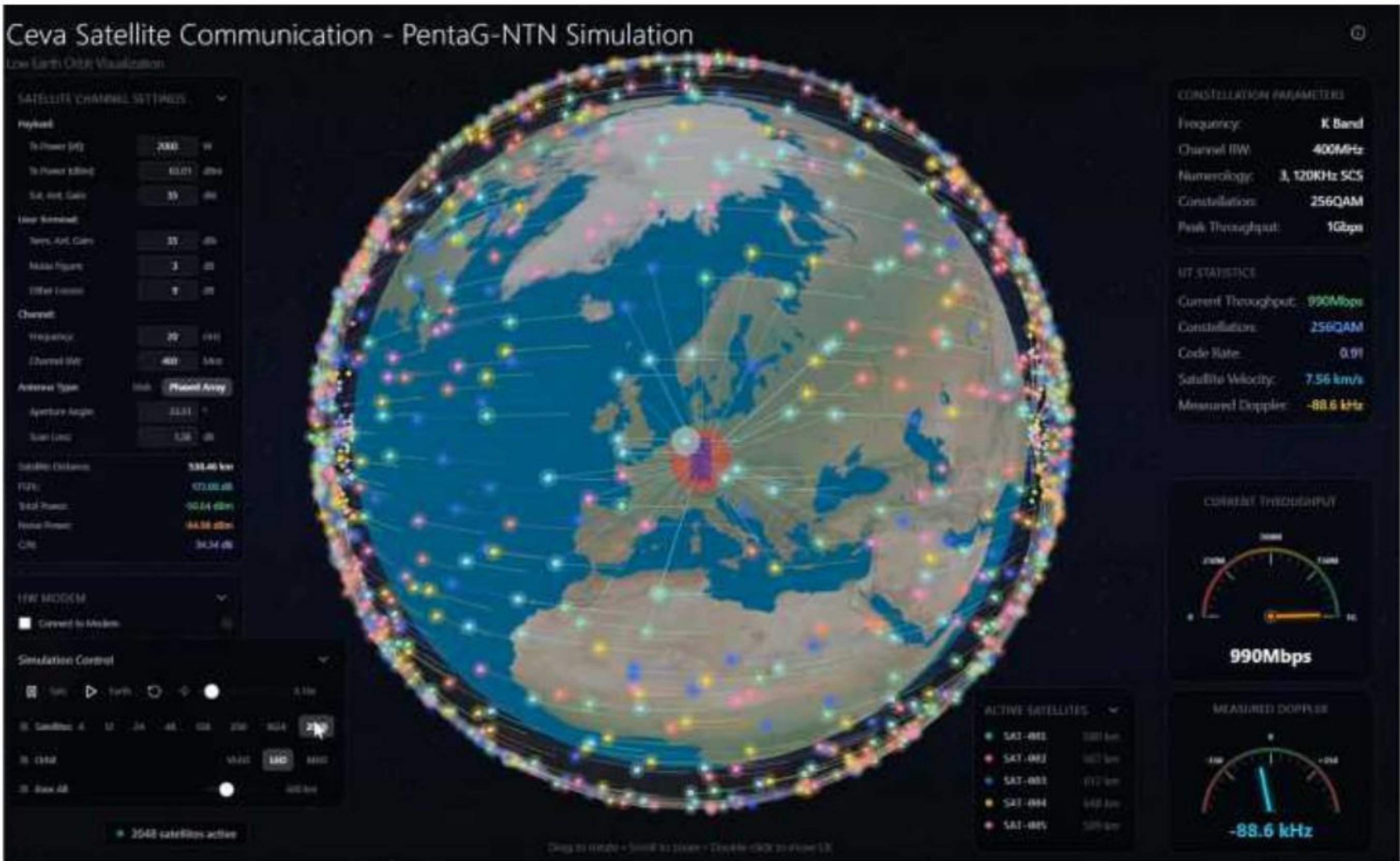
Another way to look at this is that the GPU has been engineered to be as general-purpose as possible, but this translates into being a 'jack of all trades and master of none'. As a result, performing inference using a GPU can consume significantly more power than one might hope.

This is where ElastixAI (www.elastix.ai) comes in. Their approach is based on hardware tailored specifically for AI inference workloads. That means leveraging field-programmable gate arrays (FPGAs) configured to implement the precise dataflows required by large language models (LLMs) and other AIs.

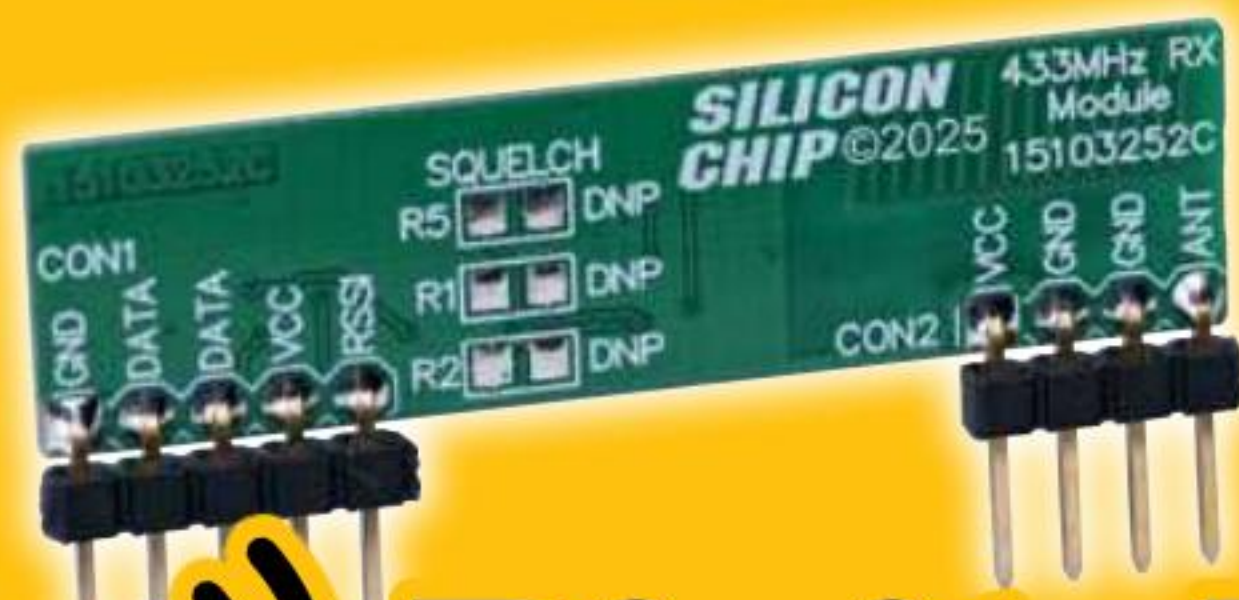
By mapping the computation more directly onto the hardware and avoiding much of the overhead associated with more general-purpose architectures, these FPGA-based solutions can deliver significantly higher performance per watt than GPUs.

This is especially true for techniques such as mixture-of-experts (MoE) models, in which only a subset of the network is active at any given time. In a world where every watt counts, that's the sort of idea that tends to attract one's attention.

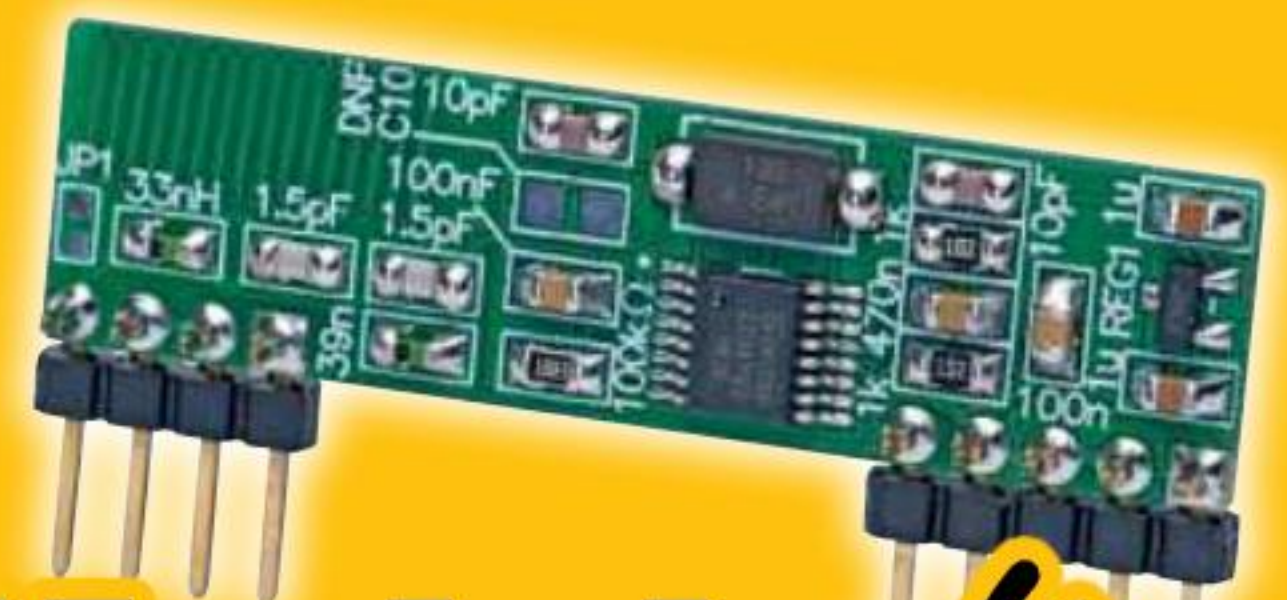
PE



Constructional Project



Tim Blythman's
433MHz



))) Digital Receiver Module (((

We recently published our version of the ubiquitous 433.9MHz LPD433 transmitter module, which performs better than many prebuilt versions. Having found a suitable receiver IC, we then created our own version of a matching receiver module, and it is better too!

We wrote about our 433MHz Transmitter Module in the May issue. That article discussed the LPD433 (Low Interference Potential Devices) RF band, which covers 433.05MHz to 434.79MHz and can be used without a paid licence.

There are some simple provisions, including that the EIRP (equivalent isotropically radiated power) must not exceed 10mW. This is of concern for a transmitter, and we explained how our Transmitter Module was compliant. Of course, this should not be a problem with a receiver, so we don't need to worry about that aspect for this project.

These sorts of receivers and transmitters are typically used to send digital data at a low bit-rate (up to around 10kbit/s) to provide a wireless link over distances up to 100m, such as around a home.

Typical applications include remote control of devices like garage doors and gates, or for sending data from remote sensors back to a base unit, as might be found in a wireless weather station.

Fig.1 shows a block diagram of such a system. As we mentioned in the earlier article, multiple layers of encoding are often used to make the best use of the medium and to allow systems to coexist with others nearby by providing identity and data validation (checksum) features.

We noted how the receivers use AGC (automatic gain control) to receive data at differing signal strengths. Simple OOK (on-off keying) means that it is quite straightforward to extract a digital signal from the ambient background RF noise.

Effectively, the receiver keeps track

of the average RF signal strength (over millisecond time scales). If the instantaneous RF signal is stronger than the average, a logical high is sent to the data output, while a low output is when the RF signal is weaker than average. (That explains why you get noise from the output when there's no RF signal; atmospheric noise on that frequency will constantly vary above and below its average.)

We've used these sorts of transmitters and receivers in numerous projects. The most recent was the Battery Powered Model Train from the December 2025 issue. John Clarke even designed a 433MHz Wireless Data Range Extender, which was published in the January 2023 issue.

433MHz receiver

Our earlier 433MHz Transmitter Module is a drop-in replacement for many commercial transmitter modules. It has the same pinout, general size and shape as those modules. The Transmitter Module article has comparative performance tests between our unit and a commercial equivalent.

This Receiver Module is intended to be a substitute for the corre-

sponding receivers. If you compare the photos above and below, you can see that we have aimed for the same pinout and size, but you'll note that our unit has an extra pin (for RSSI, a useful extra feature).

The Transmitter Module uses a Microchip MICRF113 IC and the Receiver Module uses a MICRF220 IC. These are both intended for use in these sorts of applications on LIPD bands, which makes our design task easier.

The MICRF220 IC

One of the quirks of these receiver modules is that when there is no nearby transmission, the DATA pin will produce a stream of random data. This is analogous to an AM radio (for those that remember radios before they became digital!) playing static when tuned between stations.

This can be a source of frustration for those using these modules for the first time, since the decoder (typically the right-hand microcontroller in Fig.1) must be able to separate this background noise from valid data.

One simple strategy is to look for signal transitions occurring more frequently than anticipated for the

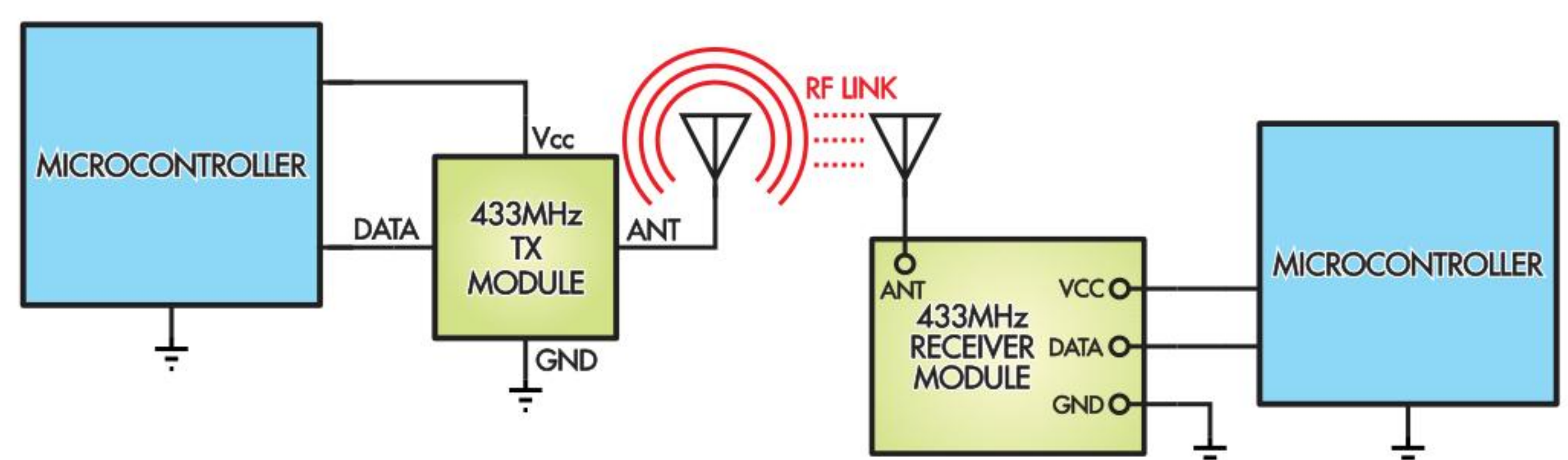


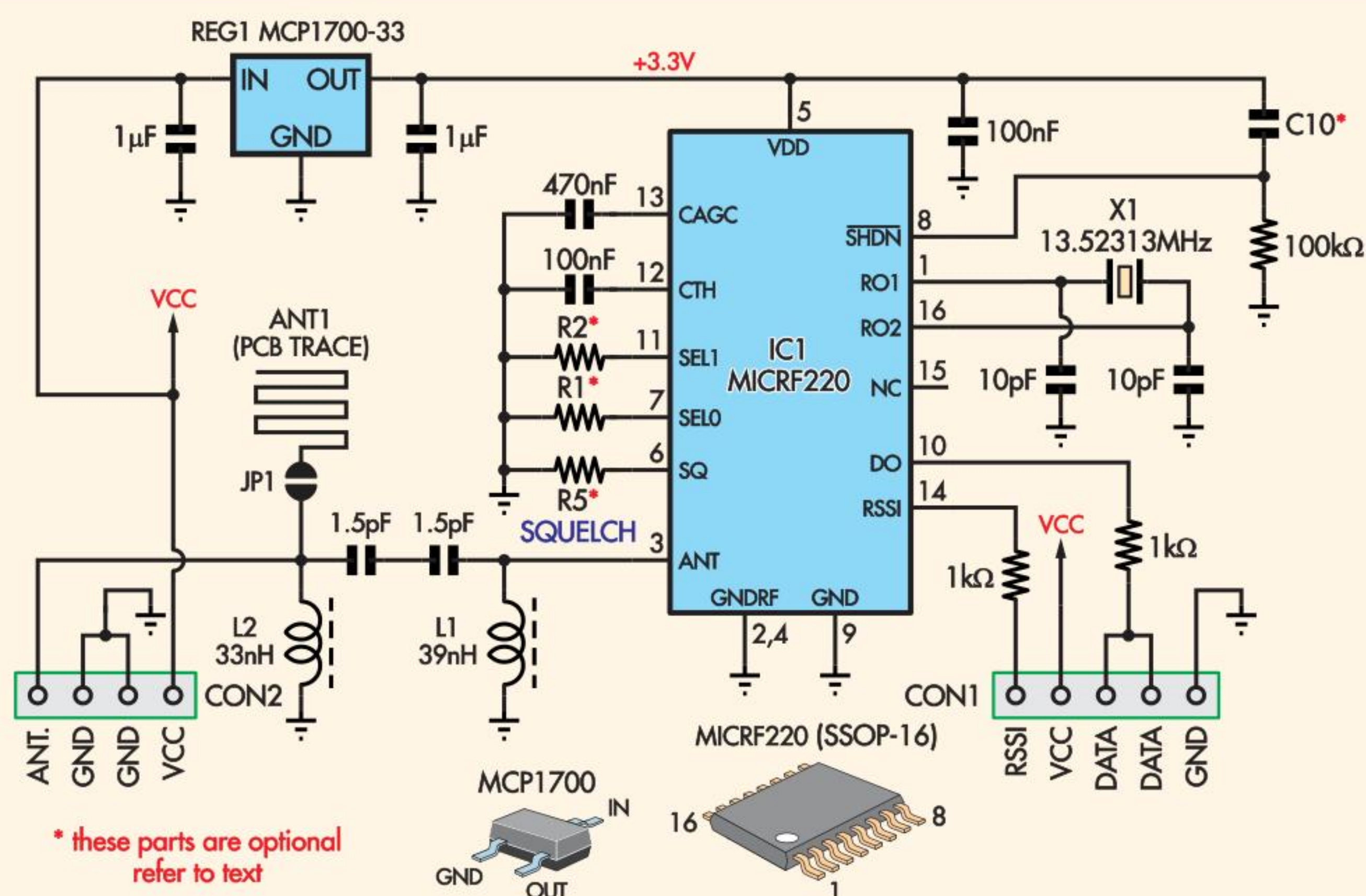
Fig.1: a matching transmitter and receiver pair form a one-way wireless link to transmit small quantities of digital data. A previous article covered the construction of a transmitter module.

433MHz Receiver Module

Features & Specifications

- » Drop-in replacement for many 433MHz receiver modules
- » Operates from 3.3-5V
- » Optional RSSI output (analog voltage, 0.5-2V)
- » 6mA nominal operating current
- » Optional squelch feature, configurable by resistors
- » Faster rise and fall times, less latency than some prebuilt receiver modules
- » Sensitivity with onboard antenna is superior to other modules with an external wire antenna
- » Short AGC settling time

Fig.2: our circuit is based on the MICRF220 RF chip for receiving digital data on the 433MHz band. It requires a 3.3V supply, provided by REG1. As well as receiving data, it provides an RSSI signal so the controller can determine whether an RF transmission is occurring and how strong the signal is.



SC
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433-434MHZ RF DIGITAL RECEIVER MODULE

expected data stream; the presence of high frequency components is typical of the white noise that occurs with no signal. So, when frequent transitions are seen, the data can be ignored.

The data sheet for the MICRF220 describes it as a “300MHz to 450MHz 3.3V ASK/OOK Receiver with RSSI and Squelch”. Squelch is a handy feature on radio receivers that can suppress the output unless a strong enough signal is received.

To maintain compatibility with the older modules, our Receiver Module can be built with or without the squelch feature. It is enabled by simply fitting a single resistor. We made this optional, since some designs may depend on the noise to detect a valid signal.

RSSI stands for ‘received signal strength indicator’, and it is exactly what it sounds like. There is an RSSI pin on the MICRF220 that produces an analog voltage related

to the received signal strength. The data sheet gives figures of 0.5V for a -110dBm RF input level and 2.0V for -50dBm. This is the extra pin on our design.

Note that the theoretical frequency range of the MICRF220 extends well beyond the 433/434MHz band. The components we have selected are intended to optimise the operation for this band; different values are needed for the likes of the 315MHz band, which sees similar use in the USA. The MICRF220 data sheet discusses this in more detail.

Circuit details

Fig.2 shows the circuit diagram of our Receiver Module. Power comes in through the various GND and V_{cc} pins on CON1 and CON2. These are chosen to match the pinouts of other receiver modules, so a few are duplicated.

The MICRF220 is a 3.3V device, so we have provided a 3.3V regulator

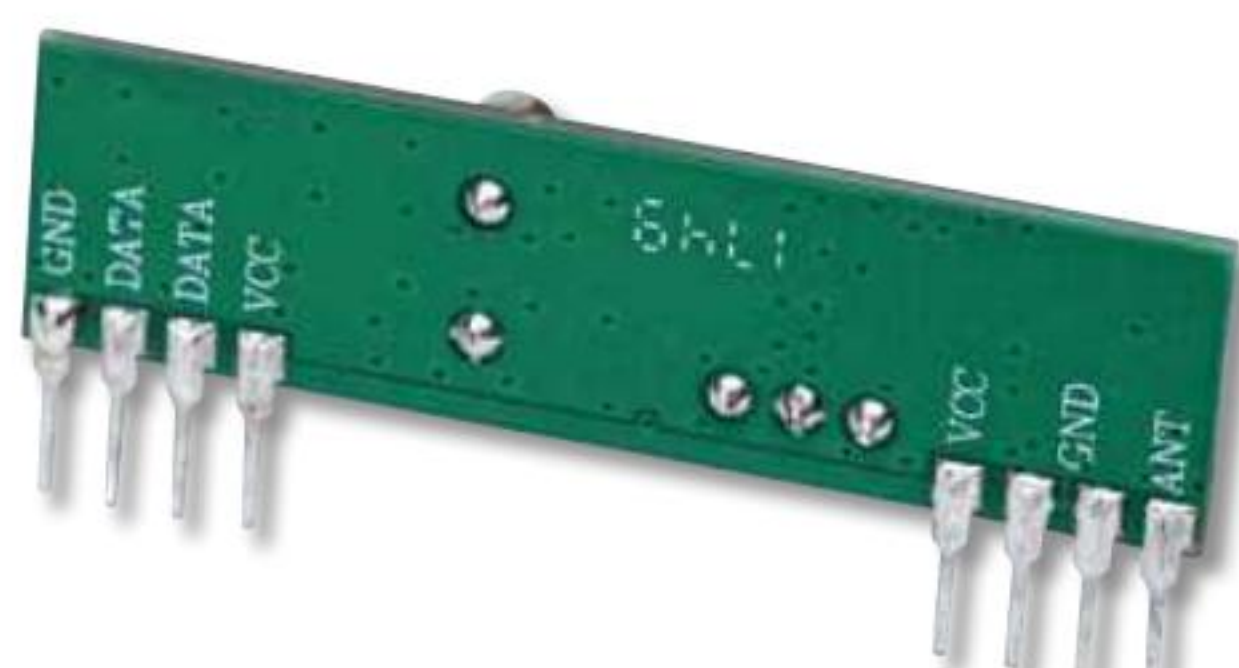
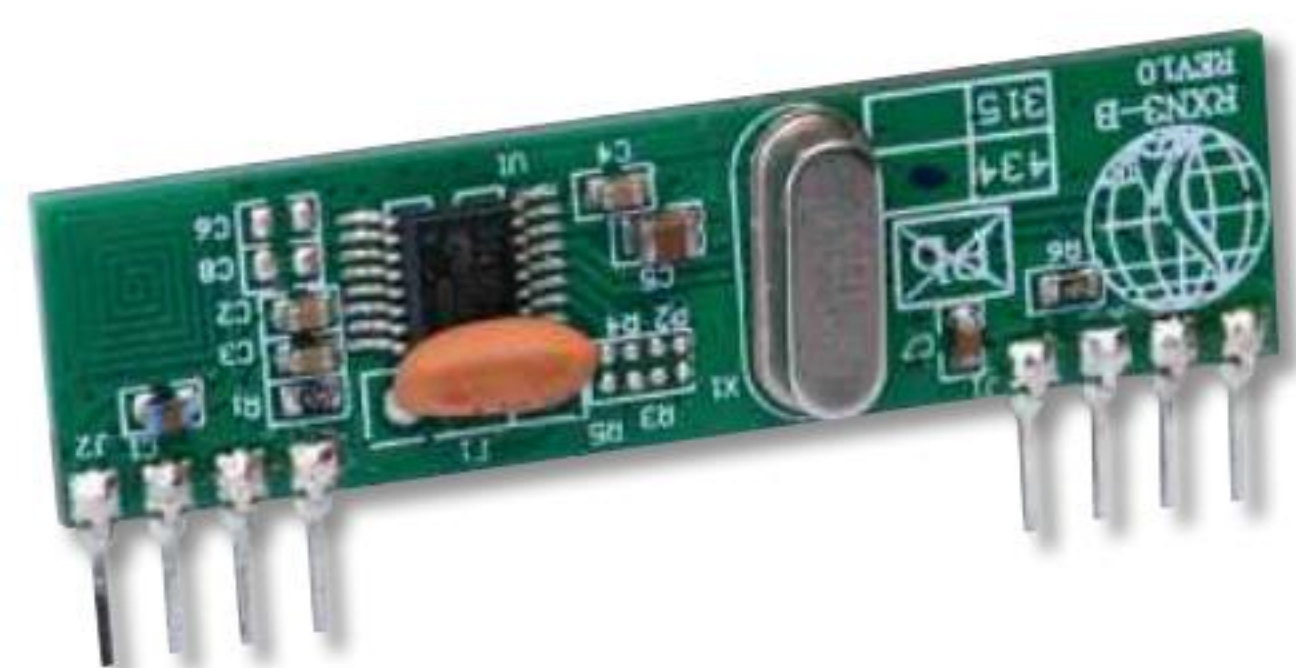
to allow operation with a 5V supply. REG1, an MCP1700, can tolerate up to 6V on its input. The two 1 μ F capacitors are recommended input and output bypassing capacitors.

The remainder of the circuit is centred on IC1, the MICRF220 receiver IC. Pin 8 ($\overline{\text{SHDN}}$) is tied low with a 100k Ω resistor to enable the chip whenever it is powered. Capacitor C10 is an optional part noted in the data sheet. When fitted, it will assert the shutdown state momentarily while the chip is powered on. We didn't find it was necessary to fit it.

Power from the regulator comes into pins 5 (power) and 9 (ground) of IC1, with a 100nF capacitor providing further bypassing. The circuit around the two inductors at lower left is the recommended matching network for the RF signal going into pin 3 of IC1 from the external antenna ('ANT') connection. Adjacent pins 2 and 4 are RF ground.

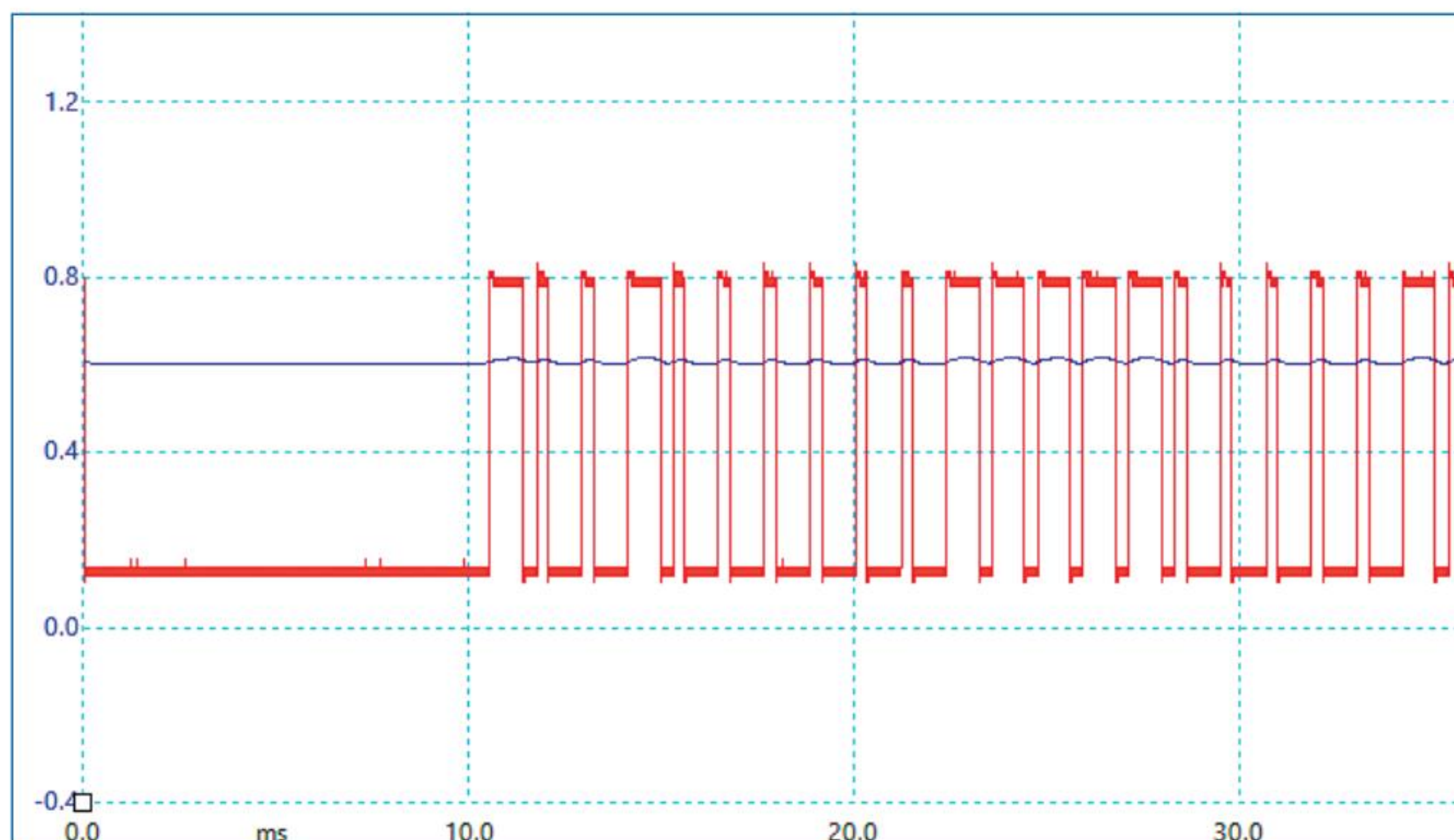
We were able to comfortably fit all the required parts in the necessary PCB area, with room to spare, so we added a PCB trace antenna. It can be connected by closing jumper JP1 with a solder blob or 0Ω resistor.

The antenna is about 16cm long, suitable for use at 433/434MHz. Adding the length of the other connected traces, it is very close to the nominal 173mm needed for a quarter-wave antenna at 433MHz. Otherwise, an external antenna can be connected via the module's ANT pin.

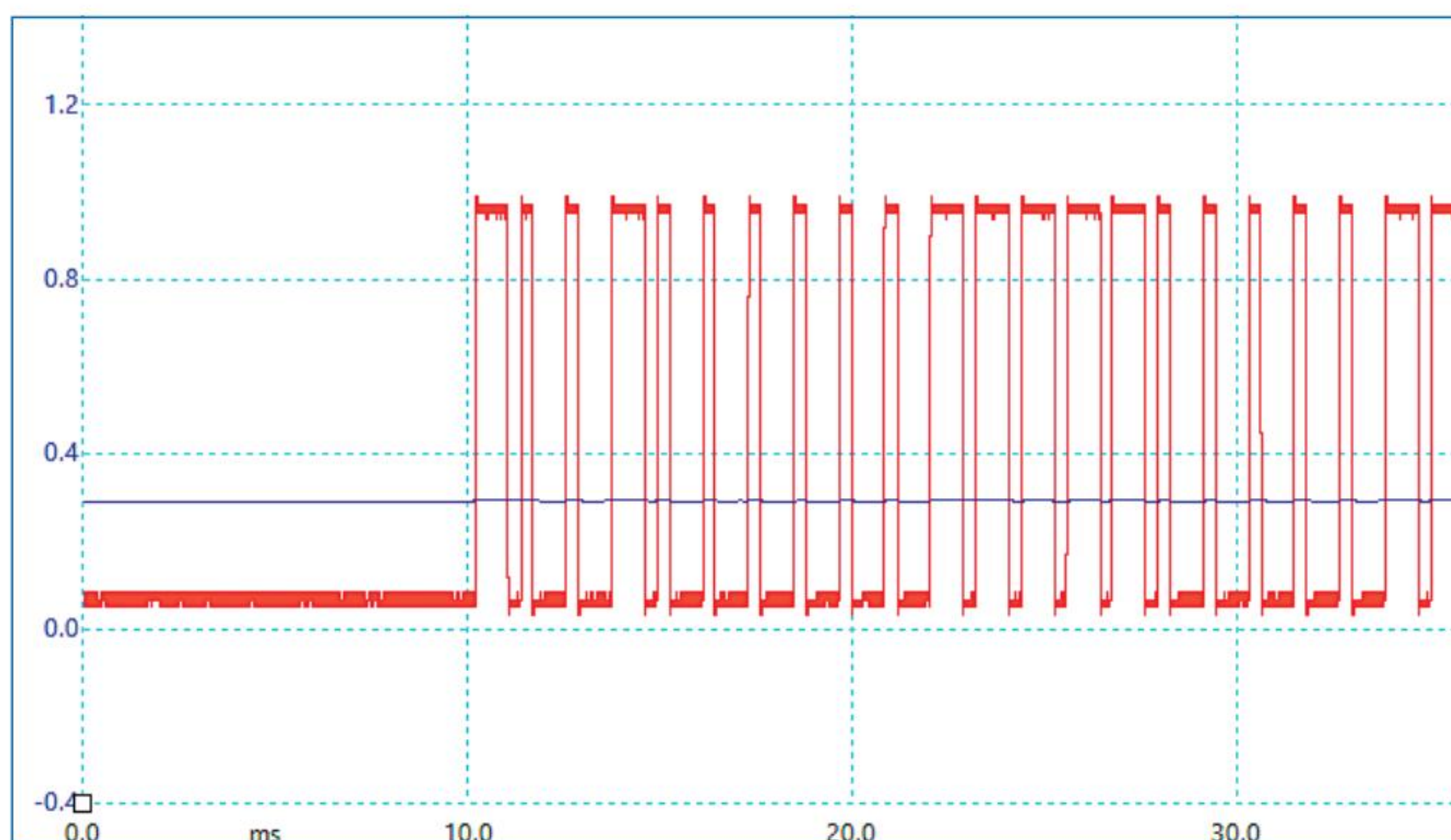


Our Receiver Module (shown in the lead photos) is the same size as boards like the commercial module shown here, but has a couple of extra features. The extra RSSI pin produces a voltage related to the received signal strength. It also has an optional onboard PCB trace antenna. These photographs are shown at 125% scale for clarity.

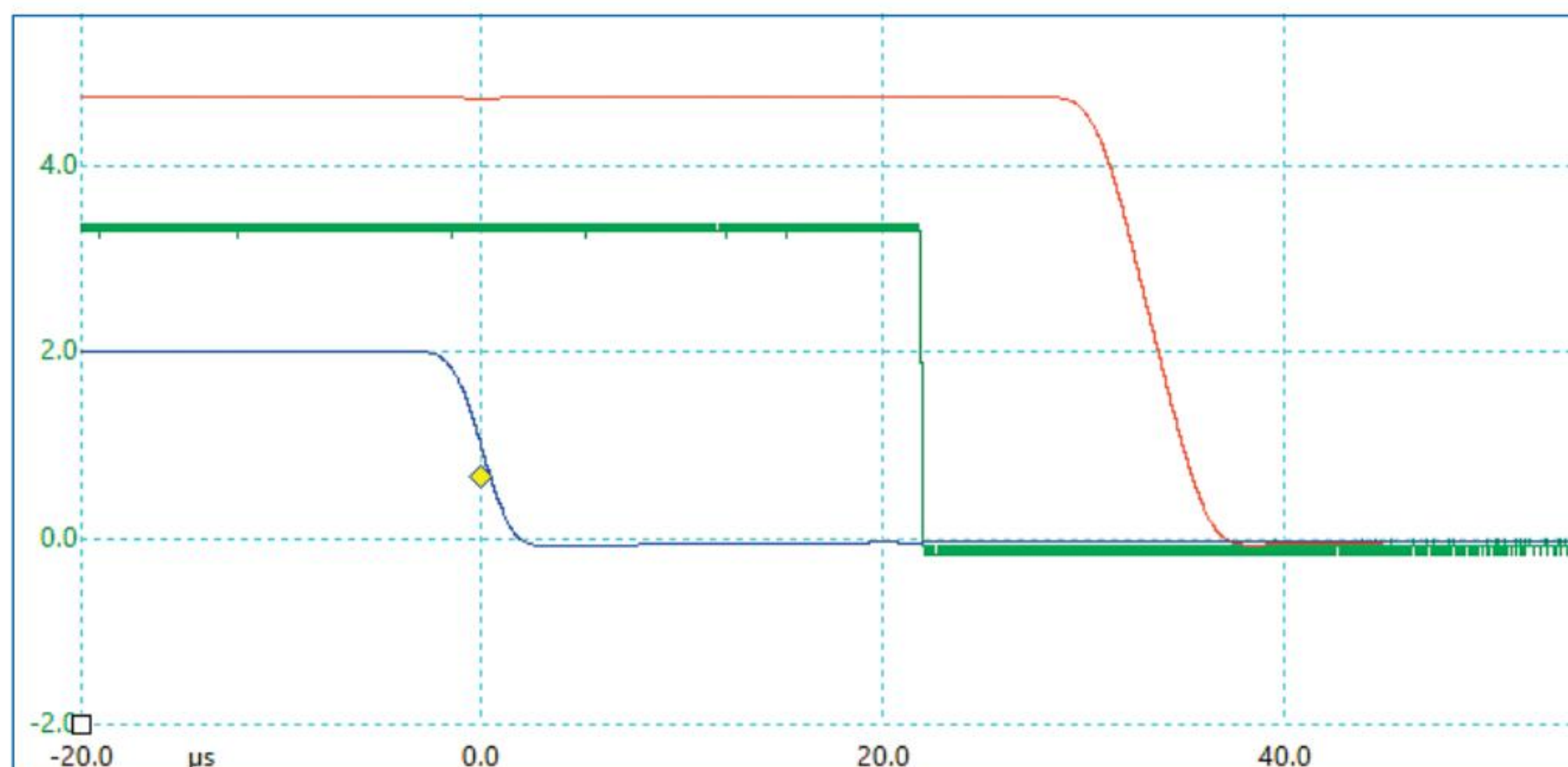
Constructional Project



Scope 1: the current consumption of our module, measured with a low-side 100Ω shunt resistor, is very close to the 6mA noted on the data sheet. It rises slightly when the data output is high.



Scope 2: the commercial module that we tested only drew 3mA during operation, although its data sheet indicates a maximum of 10mA.



Scope 3: the blue trace is a signal applied to a transmitter module, while the green trace is the DATA output from our Receiver Module. The red trace is the output from a prebuilt module. Our module is clearly quicker to respond, with sharper edges.

Pins 1 and 16 of IC1 connect to a 13.52313MHz crystal and its loading capacitors. Like the Transmitter Module, this circuit uses a $\times 32$ PLL (phase-locked loop) to generate a reference frequency.

You might notice that the crystal for the Receiver Module is a different frequency to that on the Transmitter Module (13.56MHz).

That is because the MICRF220 uses an IF (intermediate frequency) demodulator. The PLL frequency is mixed with the incoming RF signal to produce a signal with a frequency about 1MHz lower. This lower-frequency signal is easier for the IC to extract the data from.

Pins 7 and 11 (SEL0 and SEL1) select the demodulator bandwidth. We have chosen the 13kHz low-pass filter setting by leaving both of these pins to be pulled high by their internal current sources. Fitting a $10k\Omega$ resistor for either or both of R1 and R2 will change this setting.

Pin 13 of IC1 (CAGC) is connected to a $470nF$ capacitor; this value is also dictated by the data sheet and the bandwidth setting described above. The level on this pin sets the gain of the internal amplifier; it is part of the AGC control loop. This capacitor value ensures that the AGC responds at the correct rate to allow the data of interest to be received.

The RSSI signal from pin 14 is internally derived from the CAGC signal by being inverted and buffered. It is fed to the extra pin on CON1 via a $1k\Omega$ resistor. This protects the chip from potential short circuits.

The capacitor on the pin 12 (CTH) provides bypassing of an internal reference voltage that is used by a comparator to generate the output on pin 10 (DO). Like RSSI, the DO output is protected by a $1k\Omega$ series resistor between it and the external DATA pin.

Pin 6 (SQ) enables the squelch feature. When left open, an internal pullup current disables squelch. Fitting R5 will pull the pin low and enable squelch. For any of pins 6, 7 or 11, the pullup is around $5\mu A$, so a resistor of $10k\Omega$ or lower will be more than sufficient to overcome the pullup.

Operation

The MICRF220 uses around 6mA when configured for 433MHz operation. At this level, the dropout voltage

433MHz Receiver Module

of the regulator is less than 100mV, so it will not have much effect on the output voltage, even if a 3.3V supply is used.

Most 5V microcontrollers we have seen will happily accept 3.3V logic levels, and the MICRF220 works down to a 3.0V supply voltage, so this Receiver Module will be suitable for 5V and 3.3V systems.

If you are considering changing the SEL0 and SEL1 settings by adding resistors R1 and/or R2, you should check the MICRF220 data sheet closely as some other parts may need to change values. You shouldn't need to do this, as the default bandwidth settings should work fine with lower data rates.

Comparisons

We thought it was important to describe the operation of our Receiver Module and the MICRF220 because it has quite an impact on the performance of the Module compared with other receiver modules. We compared our Receiver Module (using its onboard antenna) to a commercial module with a simple wire antenna.

We used our previously described 433MHz Transmitter Module as the RF source for the tests. Our first test was to confirm the operating current of the modules. The MICRF220 data sheet notes a typical current of 6mA; the MCP1700 has a quiescent current of 1.6µA, so it does not contribute significantly to the Receiver Module's consumption.

We rigged up a 5V supply and a 100Ω resistor as a low-side current measuring shunt on a breadboard. The breadboard allowed us to change between the two modules without otherwise altering the circuit.

Our Receiver came in right on 6mA, as seen in Scope 1. You can see that the current does come up slightly when the output pin is high; it reaches 6.1mA. The prebuilt module measured just under 3mA, whether its output was high or not (Scope 2). Interestingly, its data sheet notes a 10mA maximum, so there may be more variability amongst these modules.

Latency and bandwidth

For the next few tests, we rigged up the two receivers side-by-side on a breadboard, allowing them to be seen responding to the same transmissions. This shows us the latency, or delay,

between the input and output. The output of our Receiver Module is not only faster (24µs vs 28µs on rising edges and 24µs vs 34µs on falling edges), but more symmetrical and it also has sharper edges.

We monitored the output of both receivers when a 1kHz square wave was applied to the transmitter's DATA input.

Scope 3 shows the falling edge of a pulse on the DATA input, with the two receivers responses following.

We also performed some tests to see how the receivers would respond to different OOK modulation frequencies. As we changed the frequency at the DATA input of the transmitter, we watched the receiver outputs to see how well they followed the input.

Above 10kHz, the output is 90° or more behind the input for both receivers, as seen in Scope 4. You'll see that our Receiver is still delivering a signal that is closer in time to the original signal than the prebuilt

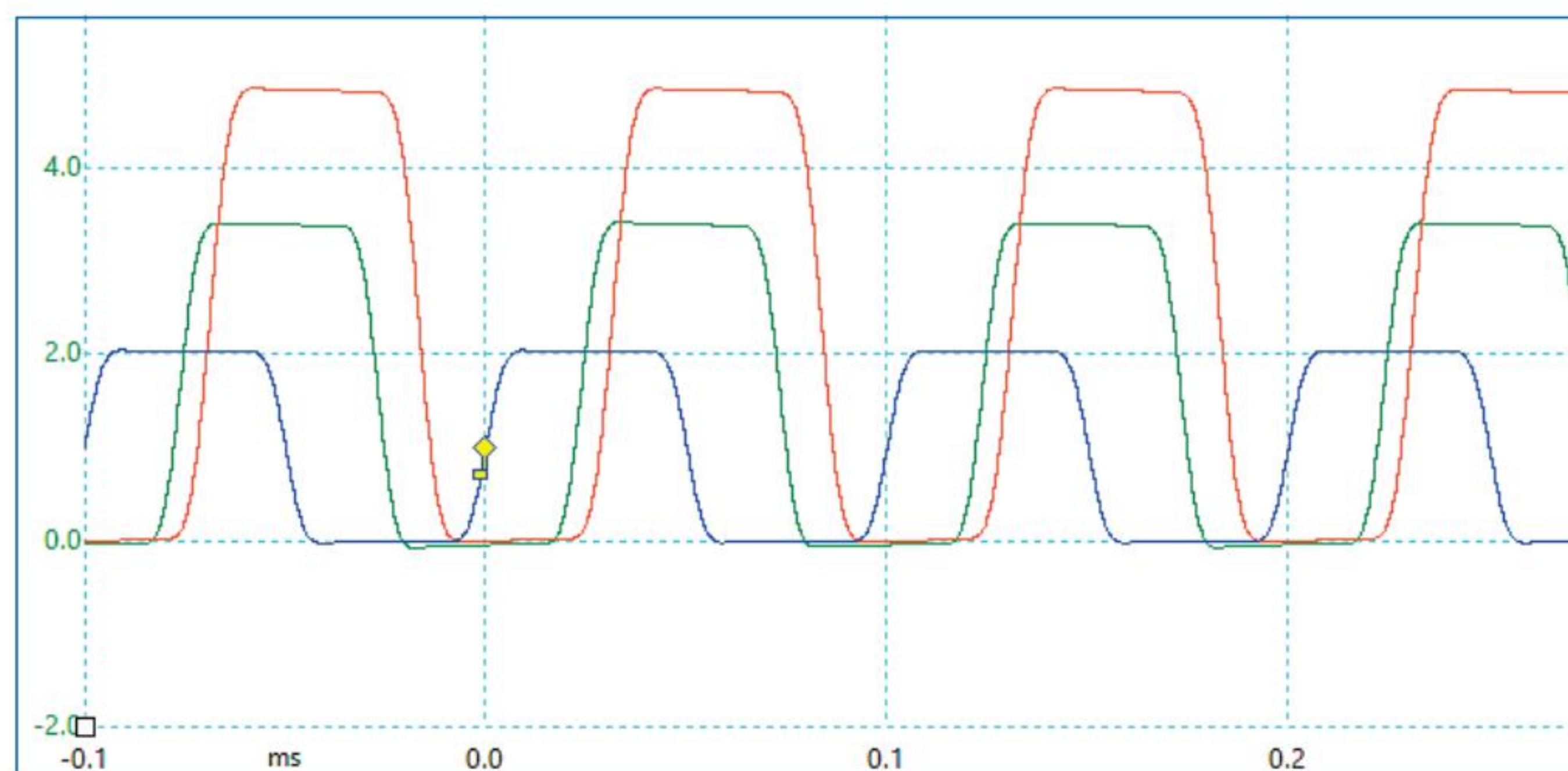
module. Both receivers are receiving a solid signal at this frequency.

Sensitivity

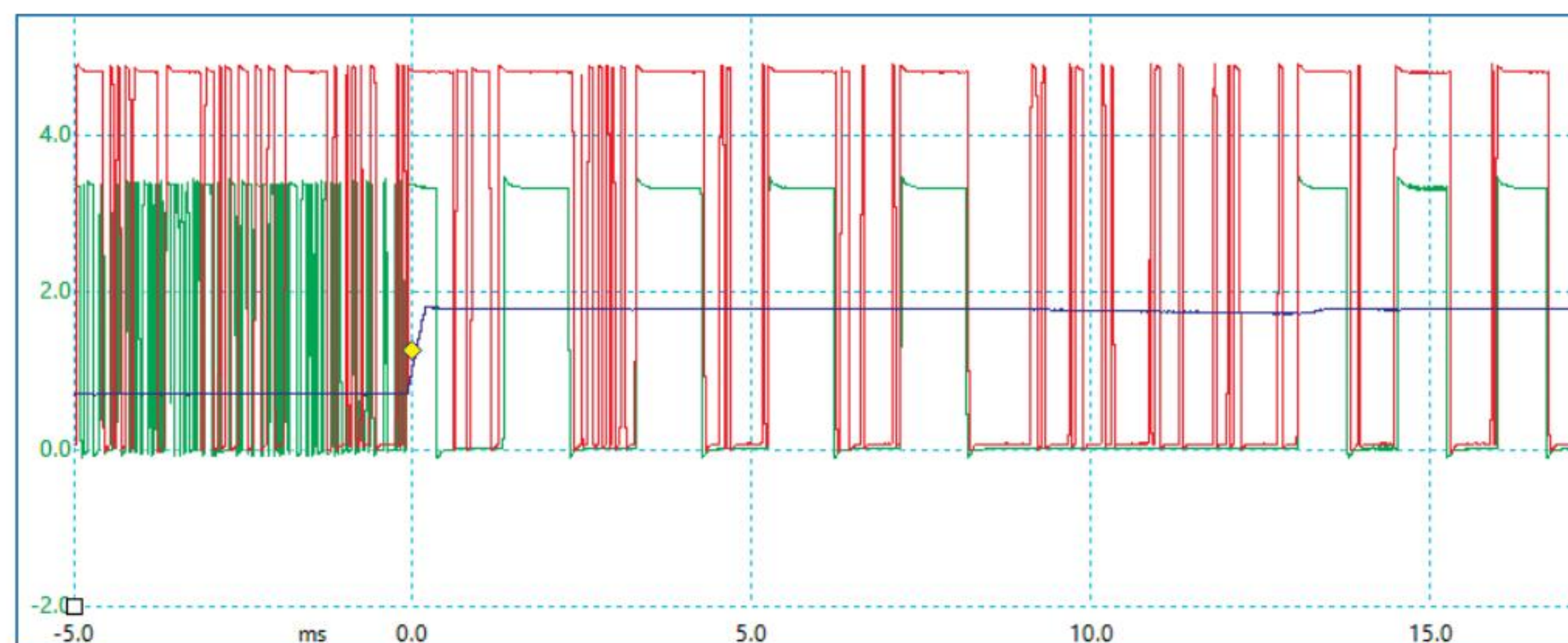
You might recall from our article on the Transmitter Module that its output power can be set by altering a single resistor value. This allows us to easily produce weak signals to compare the sensitivity of the two receivers.

We performed some tests to compare the relative sensitivity of the receivers. With the two receivers side-by-side on the same breadboard, we monitored how they responded to a transmitter on the other side of our laboratory; this was an Arduino connected to one of our Transmitters to output a typical encoded waveform.

The first test was with the Transmitter at full power, and Scope 5 shows an interesting result. Here, we see how quickly the receivers 'lock on' to the signal. Our Receiver Module settles its AGC at the correct



Scope 4: at 10kHz, a higher frequency than used for Scope 3, you can see the difference in latency between the two modules. This is quite a bit higher in frequency than most 433MHz transmissions we have seen, with 1kHz being more typical.



Scope 5: the blue trace here is our Receiver's RSSI pin, while the green trace is its DATA output; the red trace is from a commercial, prebuilt module. You can see how much more quickly our Receiver locks on to the incoming signal and starts producing valid data.

Constructional Project

Parts List – 433MHz Receiver Module

- 1 double-sided PCB coded 15103252, 11.5 × 43mm
- 1 5-way right-angle pin header (CON1)
- 1 4-way right-angle pin header (CON2)
- 1 13.52313MHz two-pin SMD crystal, 5.0 × 3.2mm (X1)
[Abracon ABM3-13.52313MHZ-10-B4Y-T]
- 1 39nH inductor, M1608/0603 size (L1) **[Murata LQG18HN39NJ00]**
- 1 33nH inductor M1608/0603 size (L2) **[Murata LQG18HN33NJ00]**

Semiconductors

- 1 MICRF220AY 300-450MHz ASK receiver IC, QSOP-16 (IC1)
- 1 MCP1700-3302 3.3V LDO linear voltage regulator, SOT-23 (REG1)

Capacitors (all M2012/0805 size, X7R 50V ceramic unless noted)

- 2 1µF
- 1 470nF
- 2 100nF
- 2 10pF NP0/C0G
- 2 1.5pF NP0/C0G

Resistors

(all M2012/0805 size, 1/8W 1%)

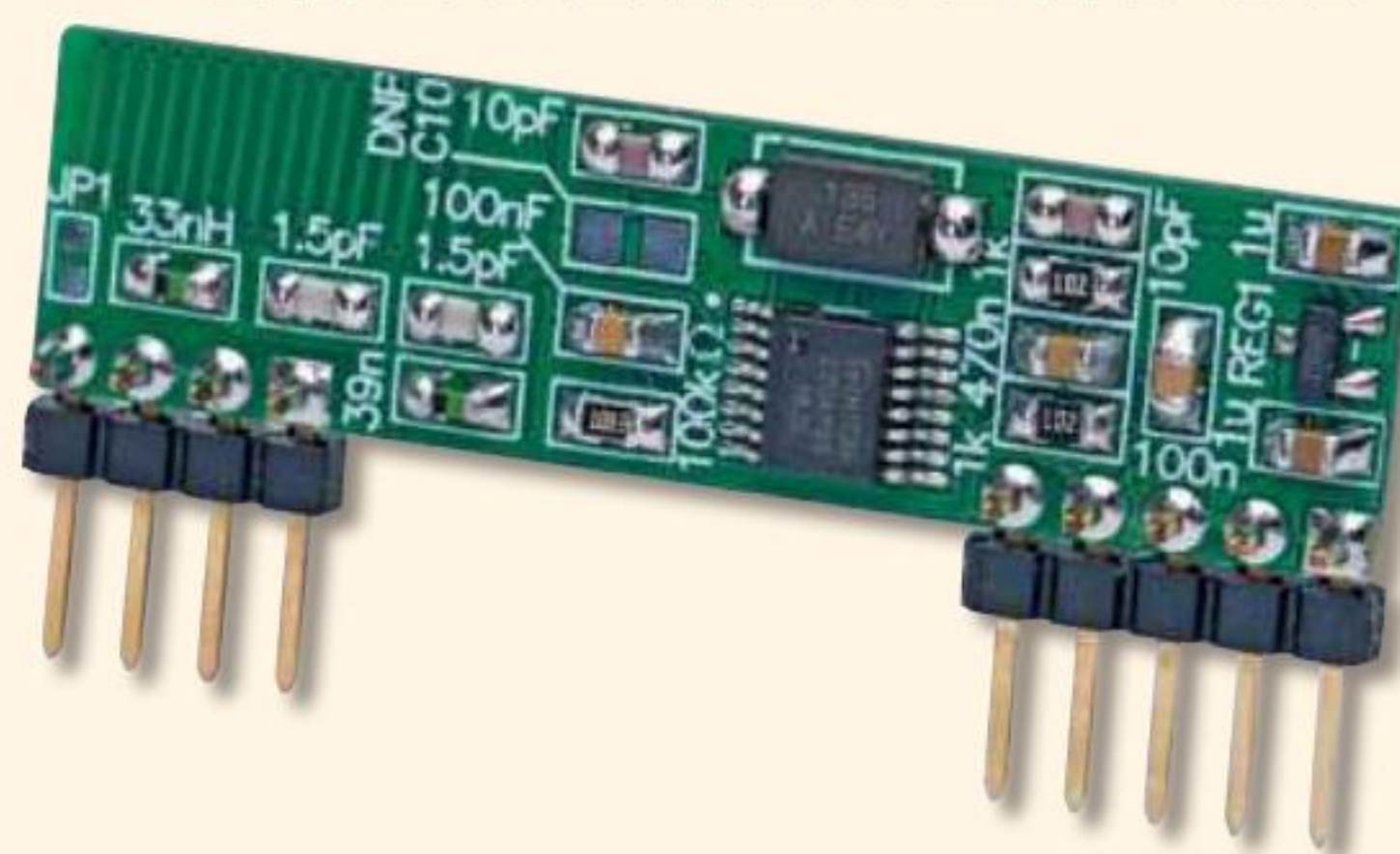
- 1 100kΩ
- 2 1kΩ

Extra resistors for option selections

- 3 10kΩ M2012/0805 1/8W
- 1 0Ω M1608/0603

OR bridge JP1 with solder

SC7447 Kit from SILICON CHIP in Australia (~£12.50 + postage): includes all the parts listed here



level a full 15ms before the commercial module; you can still see glitches in the latter's output for this time. The blue trace that we have used as a trigger is the RSSI output of our Receiver.

Then we used a 1kΩ resistor to set the output power to 12dB below nominal. Scope 6 is the result of this. The RSSI trace sits at around 1.4V or -74dBm, and our Receiver has picked out clean data, while the prebuilt module is seeing some data but is delivering glitches too. At lower levels than this, we could not see any data on either receiver.

This is useful information in that we now know that a level of around 1.4V indicates sufficient RSSI to receive a valid signal. Remember that these tests were done with the Receiver's onboard antenna; an external antenna should give even better results.

While running these tests, we also used a software-defined radio receiver to monitor the relative RSSI. It indicated that these active transmissions were only about 10dB above the background RF level.

Squelch

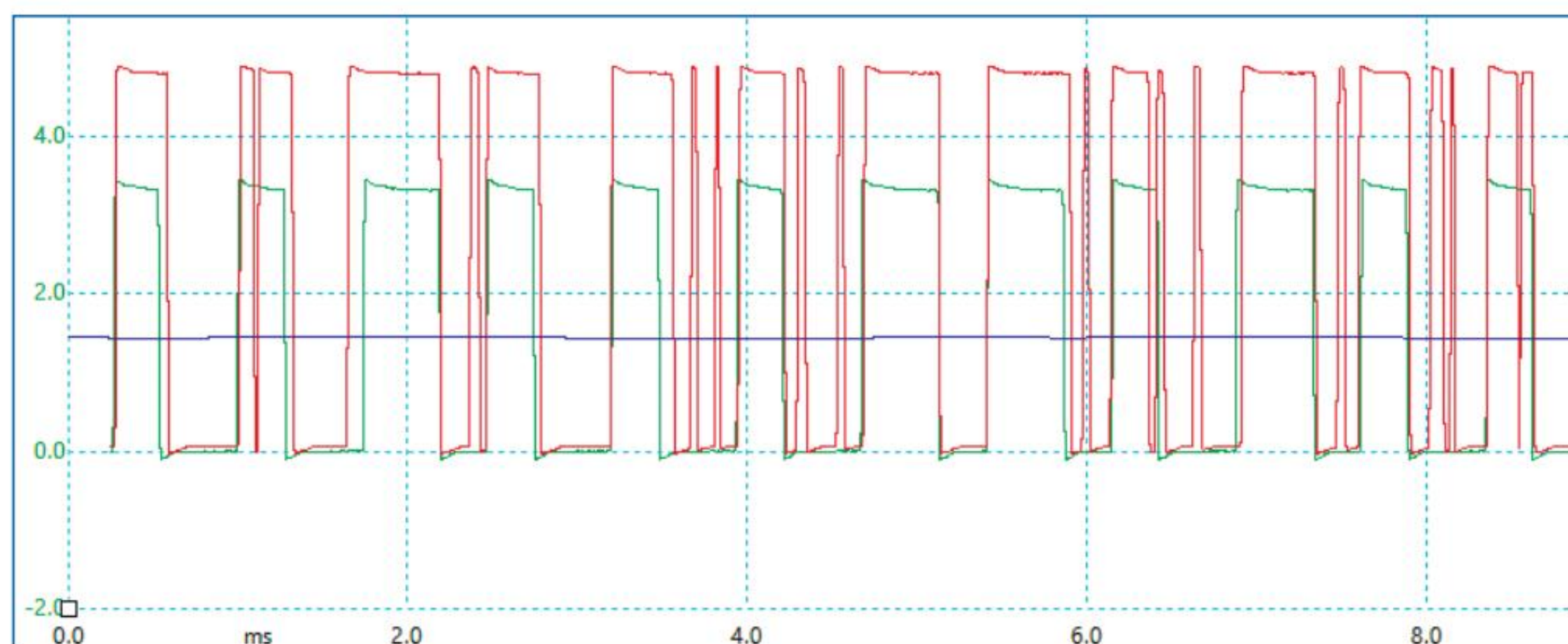
We also ran some tests to try out the squelch feature. For these, we simply shorted out the R5 pads on the PCB to pull IC1's pin 6 low. The data sheet notes the chip will "monitor incoming pulse width before allowing activity on DO pin."

So it doesn't appear that RSSI is used to control the squelch. Scope 7 shows a typical waveform with squelch active. You can see that there is still activity on the DATA line even when the RSSI is low.

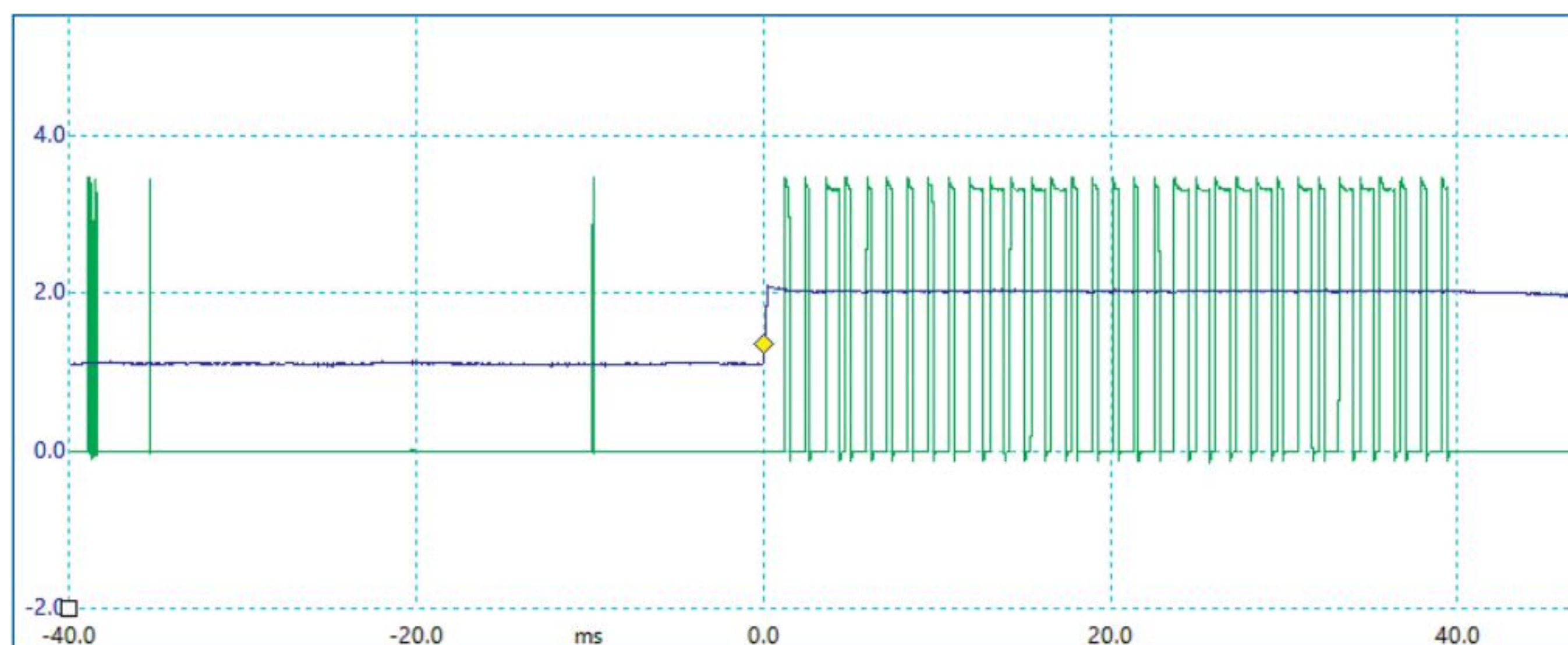
It appears that this is where the 13kHz filter is used, as signals at a higher frequency are cut off and do not appear on the output. So the squelch is helpful, but does not completely remove the need to filter out unwanted activity on the DATA pin.

Construction

The 433MHz Receiver Module uses some small SMD parts, although nothing that can't be hand-soldered with a little patience. IC1 comes in a 16-pin QSOP (quarter-size small-outline package) with a 0.635mm pin pitch, and the regulator is a SOT-23 part. Most of the passives are M2012 (0805) size at 2.0 × 1.0mm, although the two inductors are M1608 (0603) size or 1.6 × 0.8mm.



Scope 6: using the same trace colours as Scope 5, we see the two modules responding to a weaker signal. The RSSI is lower, and the prebuilt module is producing glitches that are not seen in our Receiver's output.



Scope 7: even with Squelch enabled, our Receiver still produces the occasional spurious pulse on the DATA line when the RSSI is low. So you shouldn't expect the Squelch to completely eliminate the need to reject noise on the DATA line, but it helps to reduce it quite a bit.

433MHz Receiver Module

We've used M2012-sized pads for the passives throughout to ease construction. Where possible, we have lengthened the pads on the PCB to make it easier to apply solder. This also gives a bit more room between the components.

So you'll need the standard surface-mounting gear; a fine-tipped soldering iron and some flux paste are the bare minimum. You should also have tweezers, a magnifier, solder-wicking braid and some good illumination. Your flux will probably also require a solvent for cleanup, although we find that isopropyl alcohol is a good generic option.

The Receiver Module is built on a double-sided PCB that's coded 15103252 and measures 11.5 × 43mm. Figs.3 & 4 are the overlay diagrams that show where the parts are placed. You can also refer to the adjacent photos during construction.

All the mandatory components are on one side of the PCB, as shown in Fig.3. Apply flux paste to the pads for all the components on that side. Start by placing the IC over its pads on the PCB, noting the orientation of the pin 1 marker; our chip had a moulded divot, which was easy to find.

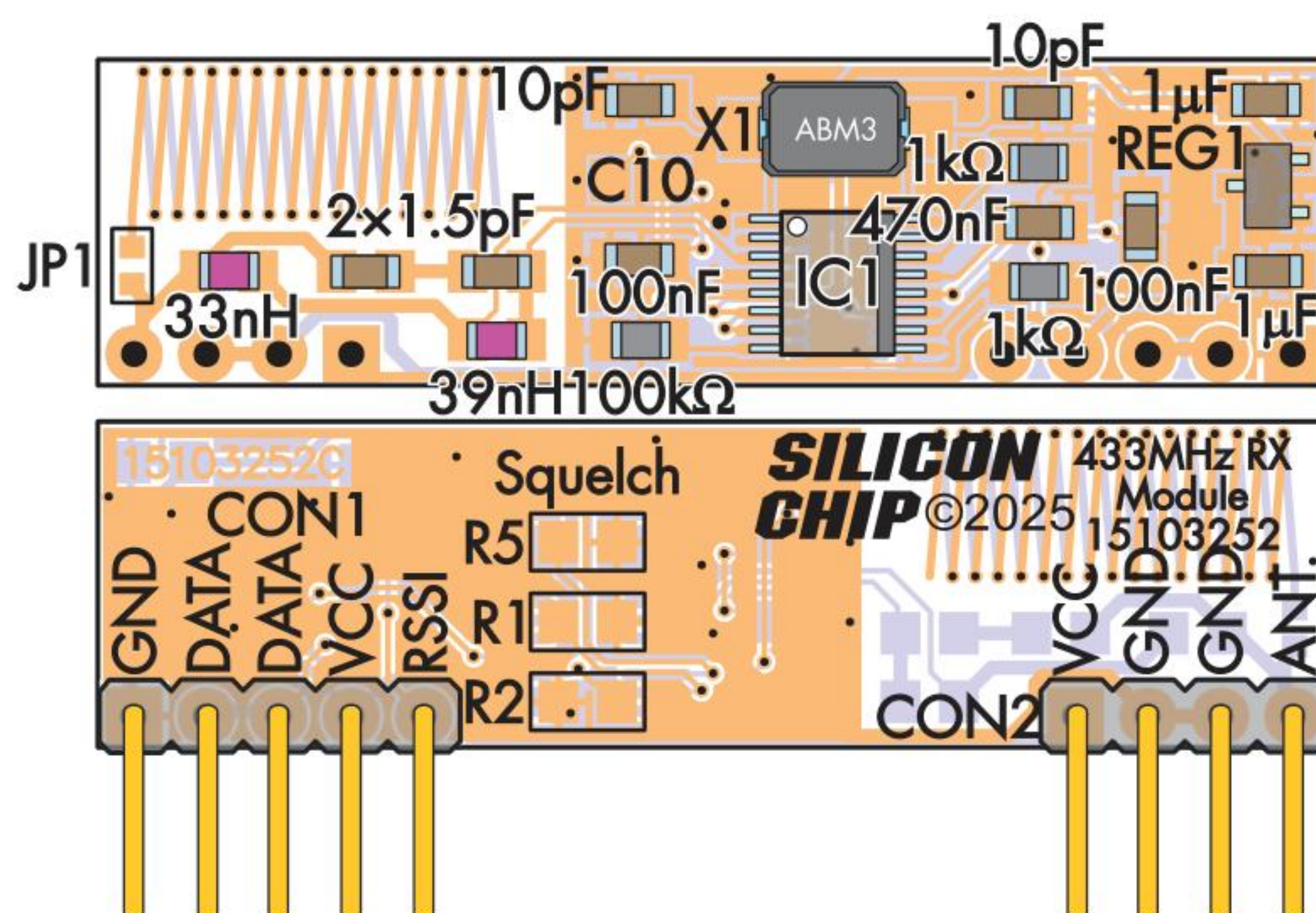
Clean the iron's tip and apply a little fresh solder. Tack one lead and check that the other pins are lined up on both sides. If so, carefully solder the remaining pins, cleaning the tip and adding extra solder as needed. Otherwise, use the iron to melt the solder and tweak the chip with tweezers until it is located correctly.

If you end up with a solder bridge joining two or more pins, add extra flux paste and press the braid against the bridge with the iron, then gently draw both away once the excess solder is drawn up into the braid.

Next, fit REG1, the SOT-23 regulator. It should only fit one way, with its leads down flat on the PCB, so place it, tack one lead and check the position. If all is well, solder the other two pins. If any joins don't look great, add some extra flux and touch the iron to the pad and pin to refresh the solder.

The two inductors at bottom left should be fitted next as they are the smallest remaining parts. We've seen some SMD inductors that only have pads on the underside, which makes them a little more tricky to solder. Don't forget that most SMD parts are

Figs.3 & 4: to use the onboard PCB trace antenna, close JP1 with a blob of solder or 0Ω resistor. The rear of the PCB shows the functions of the external pins. If you wish to enable the squelch function, you can fit a 10kΩ resistor for R5. These diagrams are at 200% scale.



usually designed to be soldered by a machine!

You might be able to make out a black mark on one end of the inductors. We've fitted our prototypes with the band to the left and on the top, which you can see in the photos. We don't think it will make any difference, but we recommend you do the same. RF can be strange and we don't want to tempt the fates!

Use the same technique of soldering one lead then the other once the location has been correctly fixed. You can check that the inductors are connected to their pads by doing a continuity test; they should read well under 10Ω.

For the 33nH part, you can probe between the ANT and GND pads of CON2. For the 39nH inductor, probe between ground and the right-hand pad of the 1.5pF capacitor directly above the inductor. If either inductor reads high resistance, add more flux and try soldering each lead again.

Solder the crystal (X1) next. It probably will have leads only on its underside, but the PCB pads are generously sized, so they will be easy to press the soldering iron against. As long as there's flux paste on the pads when you place the crystal, solder should flow between the pads and crystal. Unfortunately, you can't check a crystal for continuity as you can with an inductor.

The remaining mandatory small parts are all M2012 (0805) passives, and they are marked on the PCB silkscreen. Check their values closely against the overlay diagram, since the markings are quite small. There are nine capacitors and three resistors that must be fitted. Be careful not

to get the capacitors mixed up once you remove them from their packages; they will not be marked with values.

If you wish to enable the onboard trace antenna, you need to close JP1. An M1608/0603 0Ω resistor will work, but the easiest way to do this is to generously apply solder with an iron to both of JP1's pads. The solder mask will cause the solder to bead, but if you add enough solder, you should be able to bridge the pads (you can see we did it in our photos).

You can now enable the squelch feature by adding a resistor (we suggest 10kΩ) to the R5 pads if you want. We have also labelled this with Squelch text.

At this stage, we recommend cleaning the board thoroughly with your recommended flux cleaner or another solvent.

Allow the board to dry and scrutinise it for bridges and pins not soldered to the pads below. If you see any problems, touch up the board, then clean it and allow it to dry again.

Fitting the headers

We recommended using right-angle headers since they will match the headers found on other common modules. However, you could choose straight headers if you need to mount the Receiver Module parallel to another PCB. It will depend on your planned application.

If you are connecting the Receiver Module to an existing design, use two four-way headers. Older designs will not expect a connection for the RSSI pin, so you should leave that pad unconnected (you could run a flying lead from that pad if you want to monitor the RSSI output).

Constructional Project

We recommend slotting the two groups of headers into a longer header socket to keep them aligned to the correct 0.1in pitch before soldering (see the photo below). Solder the pins in your preferred orientation, then remove them from the header socket.

If you have jumpered the onboard antenna, the external ANT pin does not need to be connected. In this case, all the connections that are usually needed (GND, DATA and V_{CC}) are at one end of the board, and can be made using a single four-way header.

Using it

Since it is a module, the usage will depend a lot on your intended project. In general, you should connect a supply of 3.3-5V to one or more of the V_{CC} pins and one or more of the GND pins. If you have not enabled the onboard antenna, an external antenna should be connected to the ANT pin.

As we noted earlier, a 173mm-long wire (including the length of the headers and traces back to the matching network) works well as a quarter-wave antenna for 433MHz. It can be curled or corkscrewed to save space if necessary.

We have found that the Receiver Module is capable of receiving nearby signals without an external antenna; you might try this for testing purposes. In general, you should have no trouble using it to replace a receiver anywhere we have specified a similar prebuilt module.

The RSSI voltage

The RSSI pin delivers an analog voltage between 0.5V and 2.0V, so a microcontroller with an analog-to-digital converter will be well-suited to monitoring the RSSI.

Our unit varied around 0.9V to 1.1V

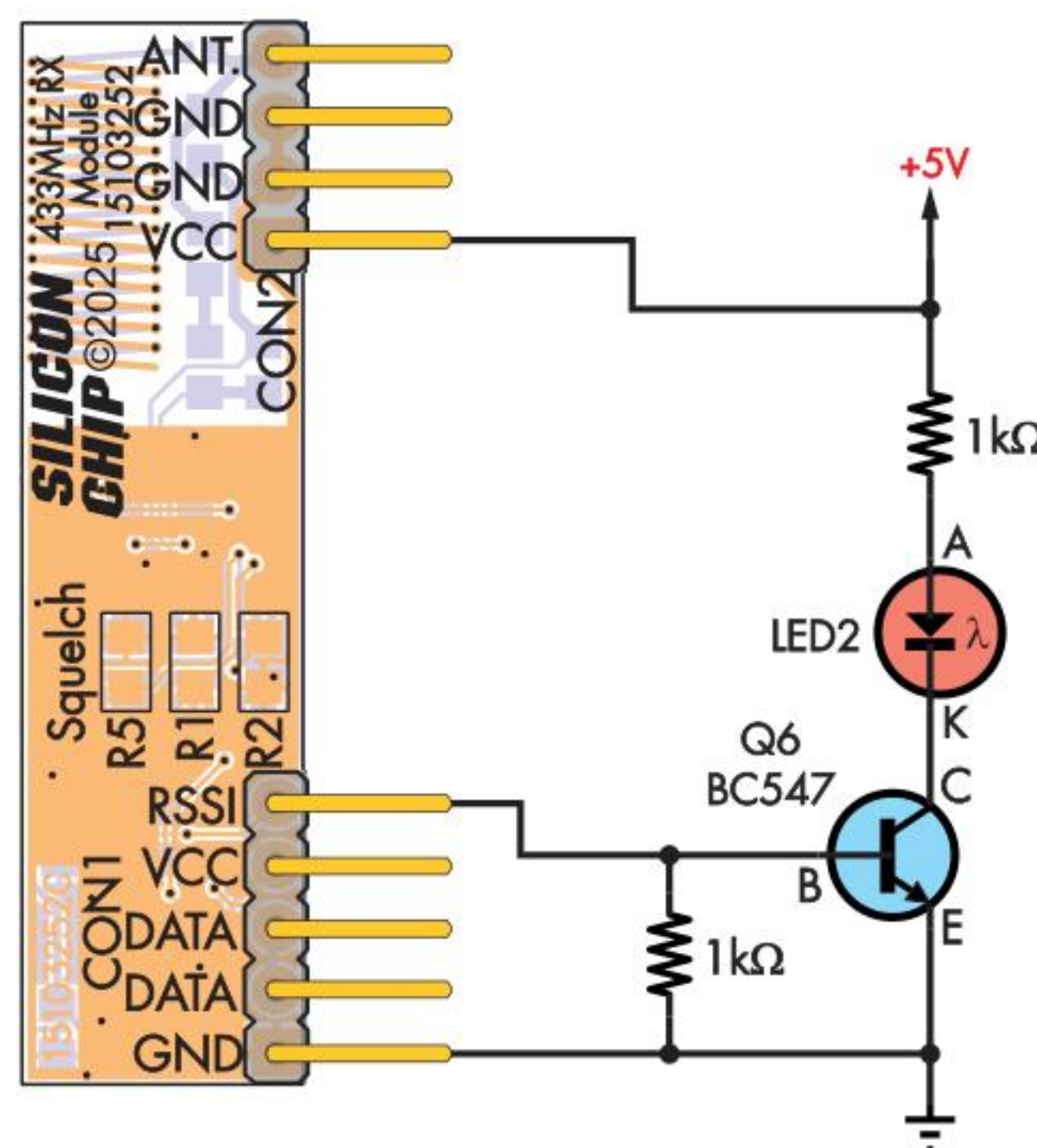


Fig.5: this simple circuit can be used to generate an indication that an RF transmission is being received based on the RSSI. The resistors set the threshold to about 1.2V, which we found to be a suitable level for distinguishing a valid signal from none.

when no intended transmission was occurring. This level might be lower in a less urban area than the location of our lab.

With an active transmitter, we saw values between 1.3V and 2.0V. This could be measured by a microcontroller's ADC (analog-to-digital converter) peripheral to detect the presence of a signal. Another option is a comparator set to an appropriate threshold.

Some micros (including the 8-bit PIC16F18146) include a comparator peripheral that could be used for this purpose. The micro could then be programmed to ignore any transitions on the DATA pin unless the RSSI indicates that a strong enough signal is present.

An NPN transistor with a 1kΩ resis-

tor between its base & emitter could also be used as a threshold detector, as shown in Fig.5. The RSSI signal is fed into the base and the 1kΩ resistor on the Receiver PCB forms a divider with the external resistor to set the threshold.

We tried this out on a breadboard and it worked quite well. You could also bypass the LED and use the voltage at the collector as an active-low digital RSSI threshold signal to a microcontroller or other circuitry.

Conclusion

Our 433MHz Receiver Module has some handy features that make it a better choice for new designs. It generally responds more quickly to an incoming RF signal.

At the same time, it is backward-compatible with older modules for use in legacy circuits that require a 433MHz receiver. The Receiver Module works with 5V and 3.3V systems, which we think will cover most cases.

The squelch feature does not appear to eliminate noisy data output during the gaps between RF transmissions, but it does reduce it. We think that the RSSI output will be more useful in testing the validity of a signal on the DATA pin.

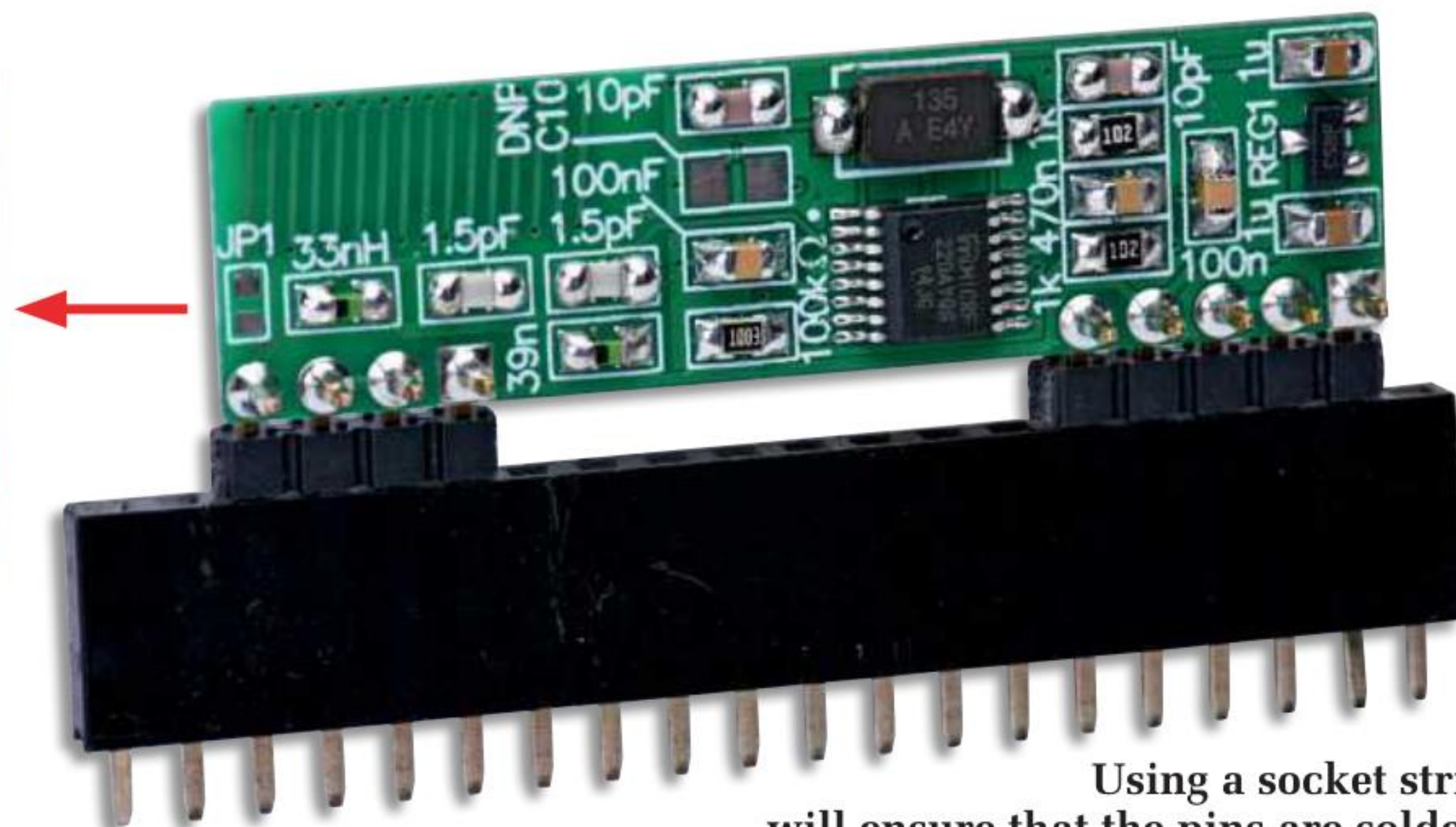
Our Receiver consumes a bit more current than the commercial equivalent, but it is still low enough that it could, for example, be powered from a microcontroller GPIO pin, allowing it to be completely powered off if necessary.

The Receiver Module is quite sensitive, even when just using the onboard PCB trace antenna, picking up all transmissions that the prebuilt module could with an external antenna. Our design still allows for an external antenna if that is preferred.

PE



We closed JP1 by bridging it with solder (you can also use a small 0Ω SMD resistor). You will need a generous amount to bridge the gap between the pads.



Using a socket strip as a guide will ensure that the pins are soldered with the correct separation even though they are in two groups.



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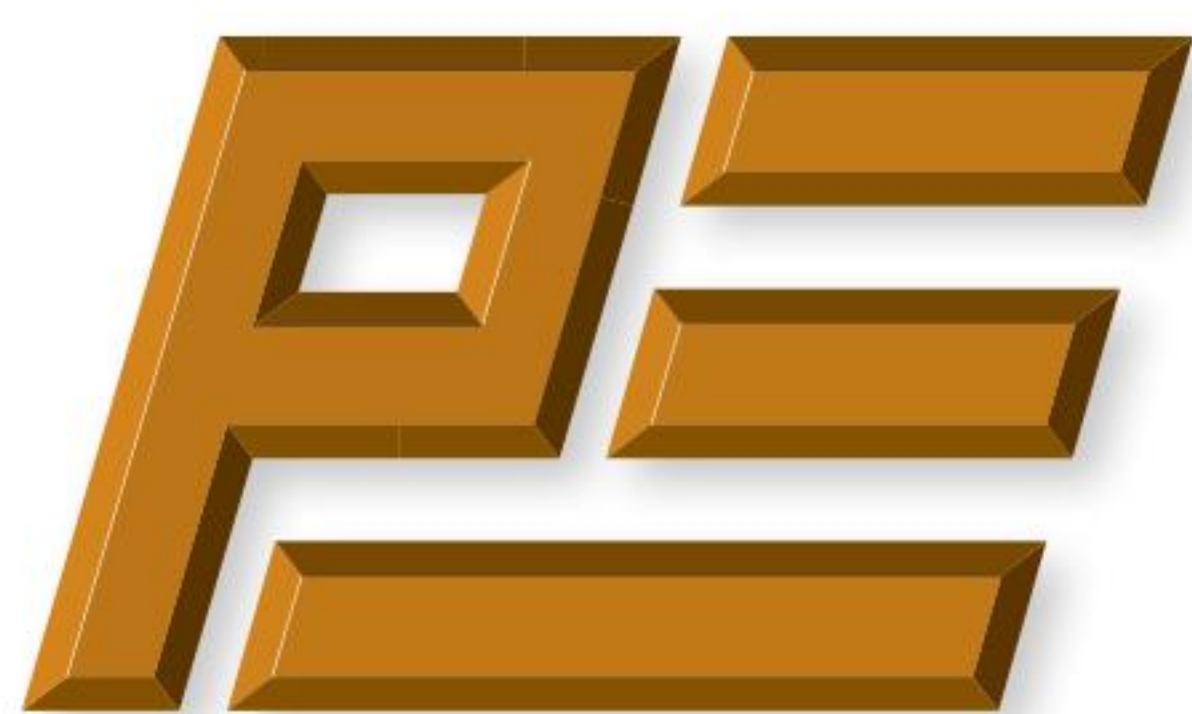
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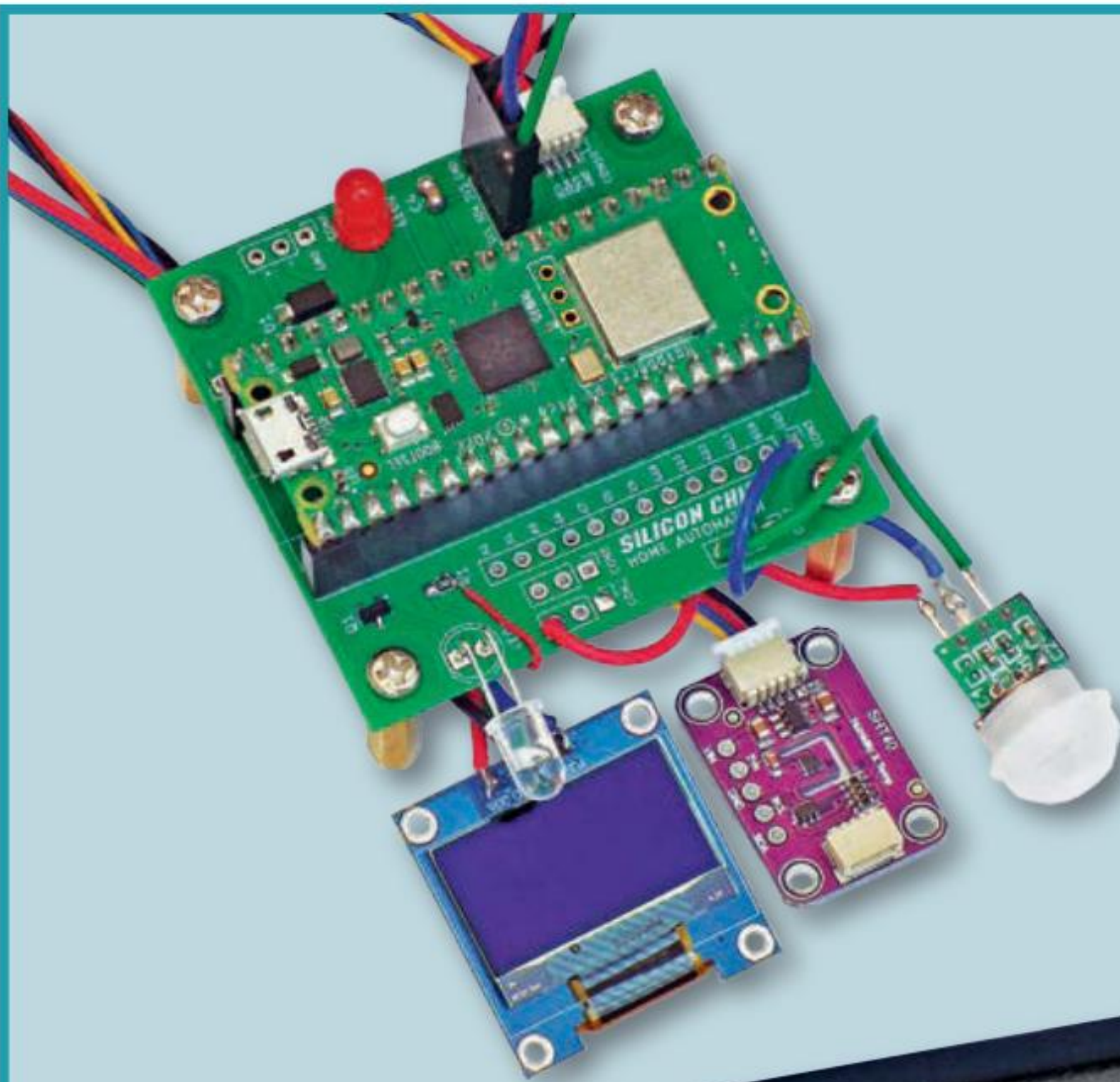
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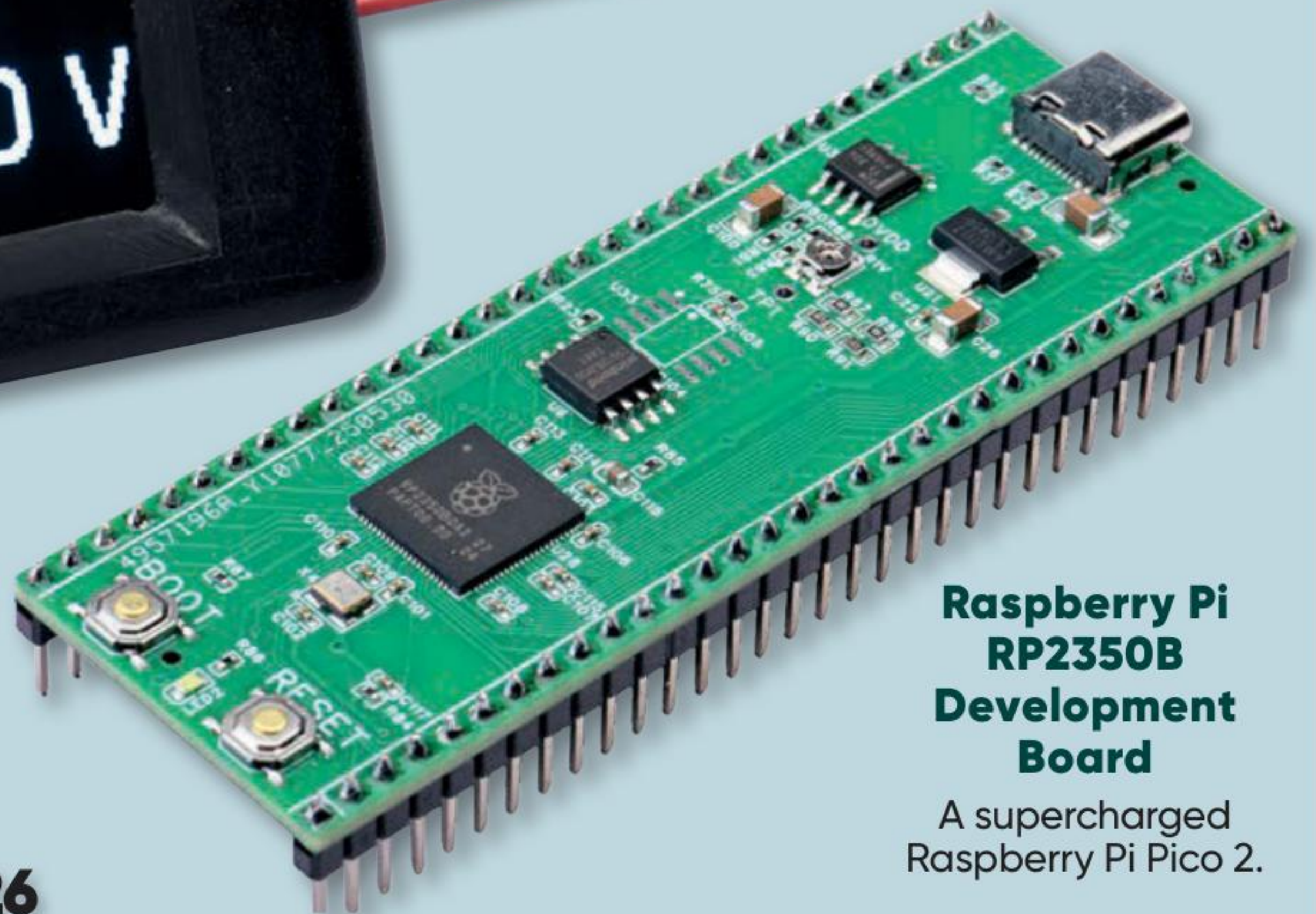
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